HUTSON

Television Receiver Theory

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# Television Receiver Theory

**PART 1**

A Textbook for Students and Technicians

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EDWARD ARNOLD (Publishers) LTD., LONDON

First published 1966

PRINTED IN GREAT BRITAIN BY
WILLIAM CLOWES AND SONS LTD,
LONDON AND BECCLES

# Preface

This book is written for technicians engaged in the servicing or manufacture of television receivers and for students of television engineering generally. It should prove particularly useful to those preparing for the Intermediate and Final examinations in Radio and Television Servicing conducted jointly by the City and Guilds of London Institute and the Radio Trades Examination Board (Course 48). In addition, sections of the book should prove valuable to students engaged in courses of study in electronic and radio engineering subjects generally, since many fundamental circuit principles are dealt with which find an application not only in television engineering, but also in the wider field of modern electronic technology.

The aim is to help the technician to gain a thorough understanding of the principles involved in the circuitry he is handling. In general, this is a two-part problem; it is necessary firstly to understand the precise purpose of each section of the complete circuit, and secondly to study methods by which the required function is achieved. At no time is the need for this more important than when equipment is working under obscure fault conditions. A clear mental picture of the purpose of a circuit, and how it is supposed to work, is surely of vital importance to anyone setting out to discover why the circuit is no longer working properly. To be truly competent, a technician must combine a sound theoretical understanding with extensive practical experience.

For several years the author has been responsible for the organisation of radio, television, telecommunication and electronic engineering courses at a technical college. Some aspects of any subject always present more difficulty to students than others. From his personal experience in teaching television engineering and general electronics, the author has singled out some of these hurdles and has tried to give perspective to them by reinforcing descriptions with explanatory diagrams. It will perhaps prove an advantage to the reader to have had some, at least, of his potential difficulties foreseen in this way.

In order to deal adequately with the subject of television receivers, covering requirements for the reception of both positively and negatively modulated signals and including a study of both valve and transistor circuitry, it has been found necessary to divide the work into two parts; the second part is in preparation. The whole is being arranged as a progressive study course and it is intended that serious students of television should progress systematically through the subject matter, chapter by chapter.

The author hopes to place in the hands of those who wish to achieve a sound understanding of television theory, a comprehensive treatment of the subject, presented in a logical and readable form. Mathematics is introduced only where necessary to achieve an adequate description and understanding of the matter being dealt with and the standard of such work is held at a very modest level so that no difficulty should be experienced here.

Where it is felt that some readers may wish to explore certain topics further, additional work is included in an appendix at the back of the book.

G. H. H.

1966

# Acknowledgments

My thanks are due to the following organisations for permission to publish extracts from circuit diagrams, technical manuals or other published literature, and also for their unfailing courtesy and assistance in answering technical queries and extending any help requested:

Alba (Radio and Television) Ltd.
British Broadcasting Corporation.
Decca Radio and Television.
E. K. Cole Ltd.
Independent Television Authority.
Mullard Ltd.
McGraw-Hill Book Company of New York.
Philips Electrical Ltd.
Pye Laboratories Ltd.
Radio & Allied Industries Ltd.
Rank-Bush-Murphy Ltd.
Thorn Electrical Industries Ltd.

In addition, I am grateful to the publishers, in particular Mr. L. C. Selwood, the technical editor, for encouragement and help in the preparation of the work and to my wife Betty for the very many patient hours she spent in typing the manuscript.

Finally my most sincere appreciation is extended to Mr. S. W. Amos, Head of the B.B.C Technical Publications Section, and to Mr. P. J. Darby, Engineer-in-Charge of the I.T.A. Transmitting Station at Dover, for their generous help and kindness in reading through the proofs and for their most helpful and constructive comments about both the technical content and its presentation.

# Contents

# Introduction

Television is a pleasingly logical subject to study. A complete television receiver looks a complicated example of electronic circuitry but those who wish to master the subject should not be deterred by this formidable appearance, even in the case of the more sophisticated colour television receiver.

As with many complex systems it is a question of 'divide and conquer'. The whole receiver is divisible into a series of sections and individual circuits, each with its own particular function. No single section or circuit is unduly difficult to understand; the complexity varies but each is logical and manageable if dissected with care. Often there are many different circuits available to achieve a given purpose. All of these cannot be included in a single textbook but, by dividing the circuits into groups and studying examples from each group, it is possible to build up sufficient experience to sort out other variations when the need arises. Development in television is continuous and, once having arrived at a good understanding of general principles and current practices, the technician should keep abreast of the subject by studying developments and innovations whenever they first appear.

In studying television it is better to begin by getting a clear mental picture of the basic idea of the whole system before embarking upon a detailed examination of individual elements. Setting up this 'mental framework' ensures an immediate grasp of the essence of the subject and quickly leads to a feeling of greater confidence. Thereafter, the details and intricacies fall into an established pattern and bring into focus something which is already partly understood and the need for which has already been discerned. The first two chapters of the book are directed towards this end.

These are followed by two chapters in which a close examination is made of television signals. Clearly it is of little use starting to study television receiver circuitry until a clear appreciation is gained of the nature of the radiated signals which the receiver is designed to accept and to process. Careful attention to the first four chapters will make the subsequent detailed study of receiver circuitry both more meaningful and more satisfying.

# 1

# Setting up a Television Picture

The aim in this chapter is to take a first look at the way in which a television picture is produced and to gain some familiarity with the general principles of a television system. The simple diagrams in Fig. 1.1 serve to illustrate the basic idea.

*Transmission of image.* At the transmitter an optical image of the scene to be transmitted is brought to a focus on a photo-sensitive *target plate* in a television camera tube. The surface of the target plate may be regarded as a *mosaic* formed by a large number of individual elemental areas. The electrical state of each minute individual area varies according to the intensity of light falling upon it at any instant. In the photo-conductive 'Vidicon' type tube it is the resistance through the target plate from front to back which varies; in other types such as 'Iconoscopes' and 'Orthicons' it is the degree of photo-electric emission from the mosaic-like surface which varies. It is not intended to examine the structure and theory of the various types of television camera tubes, but merely to establish the general idea of translating an optical image into electrical information which may then be used to modulate a radio carrier wave.

It is helpful to think of the target plate as consisting of a very large number of tiny photo-electric cells. The action of focusing an optical image upon this mosaic of cells is to produce, on the target's surface, an electrical pattern which corresponds to the picture. Since the electrical response of each minute cell represents the brightness of a corresponding minute area of the complete picture, it only remains to 'read-off' the information carried by the cells to produce an electrical waveform which represents the picture. It is obvious that the cells must be read over and over again extremely rapidly in some pre-determined logical sequence, because moving pictures are being dealt with and the light intensity falling on each area of the target is varying all the time.

The 'reading' of the cells is achieved by means of an electron beam which is made to move from cell to cell at high speed. This electron beam is brought to a focus on the surface of the target plate. In some camera tubes, the beam falls on the same side of the target plate as that on which the optical image is projected, but in others the image is focused on one side and the electron beam is directed at the other.

By means of magnetic fields set up by suitable currents in deflection coils, the beam is made to explore each part of the target plate in turn, moving over it, or *scanning* it, in a systematic pattern of lines. The target plate is held at a positive potential with respect to the cathode and the actual beam current varies according to the electrical state of the element or cell being scanned at a given instant, so producing a varying voltage across a resistor through which the current is made to pass.

The camera thus produces a voltage which changes in magnitude continuously, its actual value at any instant being determined by the intensity of the light falling upon the particular target element being scanned. This voltage, when amplified, is made to modulate a carrier wave.

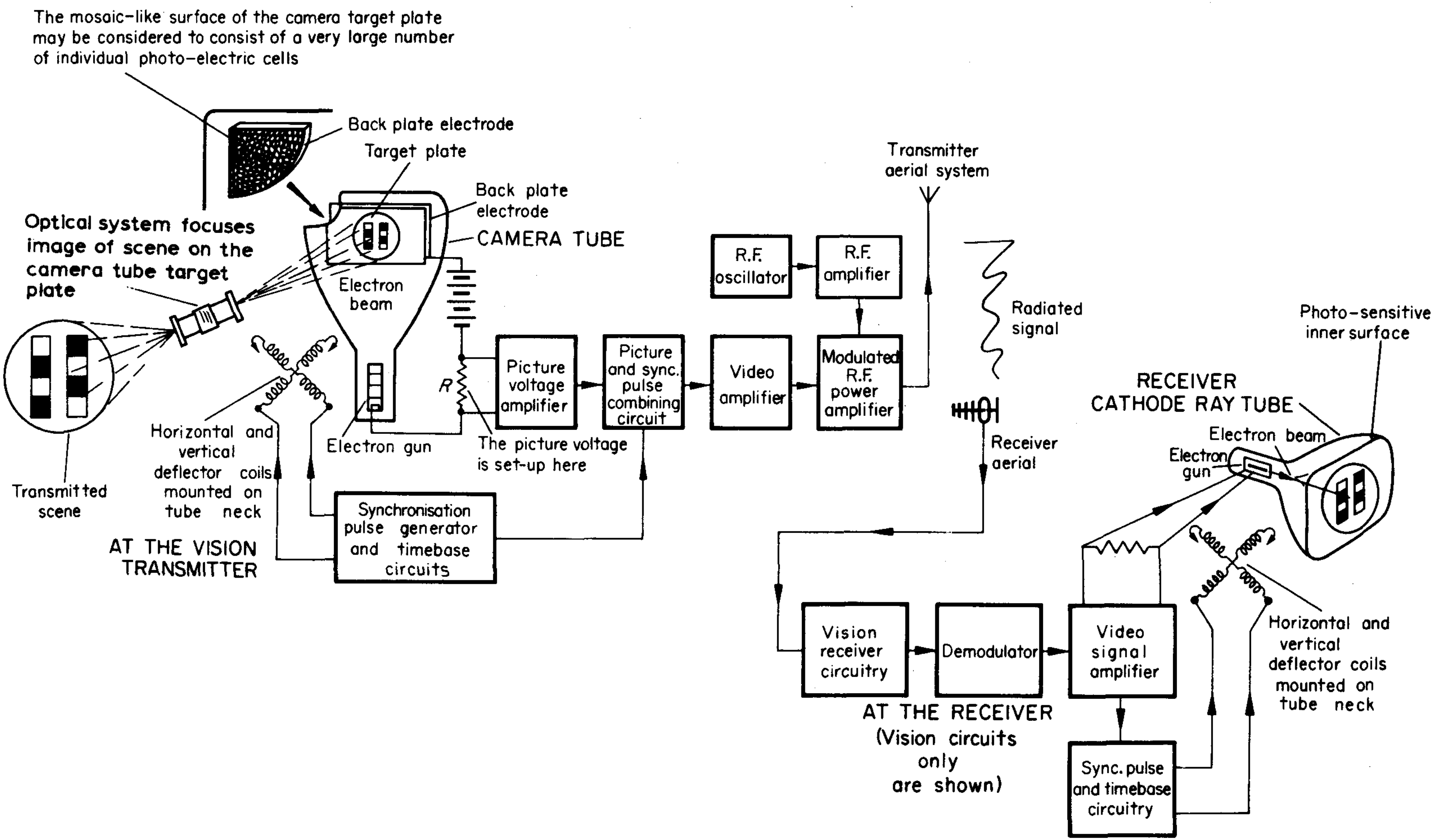



Fig. 1.1 Setting up a television picture
These simplified diagrams illustrate the basic principles of a television system. The accompanying sound channel is not shown.

*Receiving the image.* At the receiver the radio wave is demodulated and the picture voltage variation recovered. The receiver c.r.t.* screen is coated with a chemical which fluoresces when bombarded with electrons. This screen is scanned by a sharply focused electron beam which is made to move in precise synchronism with the transmitter camera tube beam. The greater the beam current the brighter the screen fluoresces. The magnitude of the beam current is controlled by the picture voltage so that the brightness changes encountered by the distant camera tube electron beam are reproduced, element by element, on the receiver screen.

The rate at which the spot of light moves is so fast that the eye is unable to follow it and so a complete picture is seen. This arises from the fact that the sensation produced, when the nerves of the retina of the eye are stimulated by incident light, does not cease immediately the light is removed but persists for a little while afterwards. The phenomenon is known as *persistence of vision* and the classic example used to illustrate it is the cinema, where a succession of stationary pictures are presented to give the illusion of continuous flowing movement.

Of course, the image on the retina does not persist for very long. If the rate at which pictures taken by the camera and subsequently projected on to a screen were slowed down, the observer would become aware first of flicker and, at slower speeds still, all movements would become jerky instead of smooth. Finally, all illusion of permanency of static scenic detail and of smooth flowing movement would be lost. The brighter the degree of illumination, the faster must pictures be presented if flicker is to be avoided.

Older films ran at 16 frames per second but the present standard is 24 frames per second. It is perhaps worth pausing here for a moment to reflect upon the necessity for synchronism of camera and projector speeds. If pictures are taken at one speed and projected at another, the apparent time in which events seen on the screen take place is altered. Thus if a film is taken at 16 frames per second, and shown on a projector running at 24 frames per second, all the events recorded are shown in two-thirds of the original time.

This is clearly apparent if a movement lasting 1 second is considered. This would be recorded on 16 frames, but if it were projected at the rate of 24 frames per second, then the 16 frames would pass through in $16/24 = 2/3$ second. This is precisely the reason why old films shown on modern projectors appear to suggest that earlier generations went about life with a great deal more gusto than their counterparts of to-day! An interesting method known as 'stretched printing' is sometimes employed to overcome this difficulty. The film is remade with every other frame printed twice. This increases the length of the film by one-third, and gives a result in which events at least take place at the correct speed.

With television, in Britain and in most other countries, 25 complete pictures per second are transmitted; this number having been chosen to give synchronism with the 50 c/s mains, with a consequent reduction of hum problems. In America, where the mains frequency is 60 c/s, the corresponding number is 30 pictures per second. It should be mentioned, however, that for Colour Television it is a disadvantage to lock the picture frequency to the mains. This is because the mains frequency varies slightly and it makes for more efficient receiver design if the picture frequency is held constant.

Finally it must be mentioned that the phosphor coating the c.r.t. screen also possesses a property of *persistence*. The screen phosphor still gives off light for a brief period after the scanning beam of electrons has passed. Phosphor persistence varies widely from one material to another, but for receiver applications the *after-glow*, i.e. the time taken for the phosphor illumination to decay to 1% of its value during bombardment, should be comparable to the periodic time, e.g. 20 ms, of one television picture.

* Cathode-ray tube.

*Presentation of the picture.* The question of flicker leads naturally to a first consideration of the interlaced raster, a raster being the pattern of lines upon which the television picture is presented. One television picture is produced when one complete set of lines has been drawn.

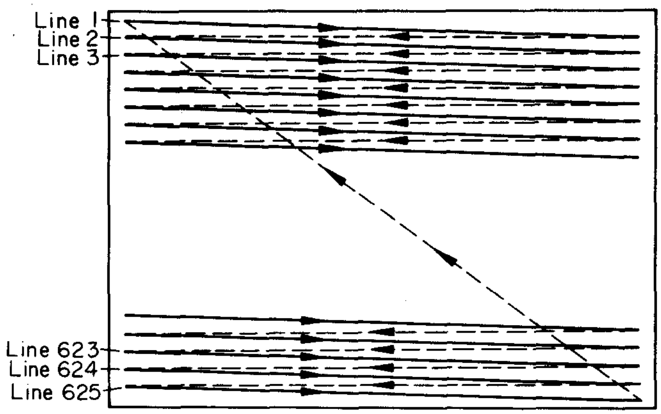



Fig. 1.2(a) Idealised 625-line sequential raster

The top and bottom of the raster are greatly extended to show the line structure. The downward deflection of the spot takes place at a uniform rate and, since the forward working stroke of each line lasts much longer than the return flyback stroke, most of the downward movement of the spot takes place during the working stroke. Thus the downward slope of the scanning lines is steeper than that of the flyback lines. The point is exaggerated in the diagram by making the flyback lines horizontal; flyback in an 'idealised' raster being assumed instantaneous.

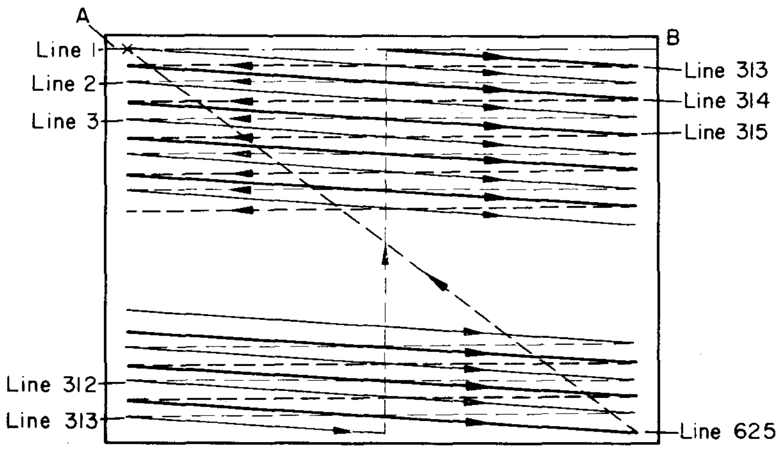



Fig. 1.2(b) Idealised 625-line interlaced raster

After $312\frac{1}{2}$ lines the vertical flyback stroke returns the spot to its starting level shown by the horizontal line AB. This *must* place the spot above the level of the centre of line 1, and subsequent lines are compelled to interlace between the first field of $312\frac{1}{2}$ lines. On completion of 625 lines the spot returns to its starting point 'X' because the vertical flyback stroke lifts it to the level AB, whilst the horizontal flyback stroke places it at the extreme left of the screen. Again, in this idealised raster, the flyback is assumed instantaneous, so that when the spot moves from the centre of line 313 to the top of the screen no time is lost, and the remaining half of line 313 is traced at the top of the screen. In practice, 20 lines are allowed for the vertical flyback after each of the two fields which go to make up one complete picture. This means that of the 625 lines only $(625-40)=585$ lines actually bear picture information.

To speak of 25 pictures per second implies, in the 625-line system, that 25 complete sets of 625 lines are drawn per second. Flicker is reduced if the number of pictures per second seen by the eye is stepped up. However, with television, the bandwidth of the vision signal goes up in proportion to the number of complete pictures radiated per second. At first sight it is tempting to think in terms of 50 complete pictures of 625 lines per second instead of only 25 pictures per second, but this would lead to a prohibitive bandwidth, as will be demonstrated in Chapter 3. By an ingenious method it is possible to arrive at the same reduction in flicker that 50 pictures per second would give, without stepping up the picture rate or the bandwidth at all. This is done by employing an 'interlaced' rather than a 'sequential' raster.

*Sequential raster.* The difference between these rasters is illustrated in Fig. 1.2. Suppose the British 625-line system is considered. A sequential raster, as suggested by the name, would be set up by drawing the 625 lines one under the other, starting with line number 1 at the top and finishing with line 625 at the bottom. For this to take place it is obvious that the vertical (i.e. the field) timebase must run at 25 cycles per second. Also, since 25 pictures per second, each of 625 lines, have been stipulated, the speed at which the line timebase must work is seen to be $25 \times 625 = 15{,}625$ c/s. Working backwards to make the point doubly clear, if the spot is deflected from left to right to draw horizontal lines at the rate of 15,625 lines per second, and is simultaneously deflected from top to bottom at the rate of 25 times per second, then the number of lines traced across the screen for each movement of the spot from top to bottom is $15{,}625 \div 25 = 625$ lines. For this purpose the return journey of the spot, from the bottom to the top of the screen, is assumed to be instantaneous.

*Interlaced raster.* In the interlaced system the 625 lines are divided into two sets of $312\frac{1}{2}$ lines. The first $312\frac{1}{2}$ lines of the complete picture are traced one after another in the same way as before, but the lines are double-spaced and line 313 is at the bottom of the screen. Half-way through line 313 the spot is returned to the top of the screen and the remaining $312\frac{1}{2}$ lines are traced interleaved between the first set. This is achieved by running the vertical timebase at 50 c/s instead of 25 c/s. The spot is therefore deflected to the bottom of the screen in half the time and, as the line timebase is running at the same speed as before, it has only completed one-half of the total of 625 lines when the spot reaches the bottom of the screen.

Working backwards as before, 15,625 lines are drawn across the screen in 1 second, during which time the vertical timebase shifts the beam 50 times from top to bottom. This means that the number of lines traced for one excursion from top to bottom is $15{,}625 \div 50 = 312\frac{1}{2}$.

The interlaced system much reduces flicker since the area of the screen is being covered at twice the rate. Of course the result is not the same as if 50 complete pictures per second were shown, because each individual line still only appears at the rate of 25 times per second. Close examination reveals that *line flicker* is visible, but this is tolerable and the over-all effect of doubling the speed of vertical movement of the spot is a great improvement and approximates to the ideal 50-complete-pictures-per-second arrangement suggested.

It should be noted that the diagrams of Fig. 1.2 show idealised rasters not obtainable in practice. This is because it has been assumed that the spot is able to move instantaneously from the bottom to the top of the screen during the flyback stroke. In practice this is not so, and the time taken for this movement is equivalent to the time duration of several strokes of the line timebase. On its way up, the beam is swept several times across the screen by the line timebase so that several of the lines are 'lost' and not available for bearing picture detail. The actual number of *active* picture lines is 585 for the British 625-line system and 377 for the 405-line system.

The reason why an odd number of lines is always employed in television systems is now

apparent. To set up an interlaced raster it is necessary to feed two regularly-spaced synchronisation pulses to the field timebase per one complete picture; i.e. pulses must arrive at the rate of 50 per second since there are 25 pictures per second. One of these pulses must arrive in the middle of a line and one at the end of a line. If an even number of lines were used, then the two sets of lines for the two successive vertical movements of the spot per picture would be bound to fall on top of one another. The subject of interlacing will be explored further in subsequent chapters.*

*Visual response to the television image.* As an example of the way in which television images

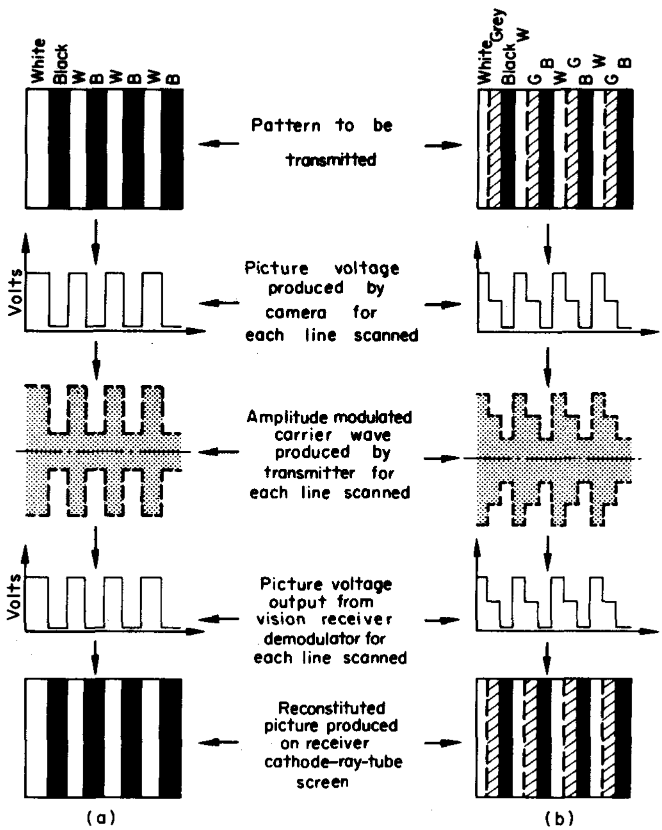



Fig. 1.3 These diagrams summarise the cycle of operations in a television system, and show how the 'shape' of the picture voltage, and of the modulated carrier wave, varies in sympathy with the brightness changes encountered along each line. (*N.B.* Positive modulation has been assumed here. The difference between positive and negative modulation is explained in Chapter 2.)

* It is necessary to draw attention to a confusion of meaning between British and American usage of the word *frame*. This word has long been used in cinematography to describe a complete picture. In British television, however, the two interlaced halves of each complete picture are often referred to as frames, and there are said to be two frames, each of 312½ lines, in one complete 625-line picture. In the U.S.A. the two halves of the interlaced picture are each called 'fields' and the word 'frame' stands for a complete picture. Since the confusion centres around the word 'frame' it has been suggested that it should be dropped altogether in both countries and the word 'field' used for each of the two interlaced halves, with the whole being called a picture. Nevertheless the word 'frame' has been widely used for years by both countries to describe vertical deflection circuitry. Currently both the words 'frame' and 'field' are being variously used to describe such circuitry. In this book the word 'field' will be used for this purpose.

are set up, suppose that the scene to be transmitted is a simple pattern consisting of a succession of eight vertical stripes coloured alternately white and black. (See Fig. 1.3.) Along each horizontal line scanned, the camera will translate this pattern into a voltage which rises sharply to maximum at the edge of each white stripe, maintains this level for the width of the stripe and falls to a minimum for the duration of each black stripe. The radiated radio wave will have an envelope of this same shape, and the output from the demodulator at the receiver will show the same voltage waveform. This, duly amplified, will be presented to the receiver cathode ray tube. The beam current will be maximum during the white stripes and cut-off during the black. Obviously this produces on the screen a series of bright white dashes followed by dashes where there is no fluorescence at all. The latter will appear jet black because of the contrast between these areas and the intense white light on each side. Each successive line scanned produces white and black dashes directly beneath those of the former lines so that the white and black striped pattern seen by the distant transmitter camera, is reproduced at the receiver. It will be appreciated that the blackness is itself an illusion since when the receiver is switched off the tube face will look white under normal room lighting conditions. There is a limit to the range of light intensities which the eye is able to accommodate at the same time. It is a question of *relative* light intensities. Compared with the brightly illuminated white columns the adjacent stripes *appear* completely black. The eye responds to this contrast of light intensities by recording bright light and darkness. The areas recorded here by the eye as being black would in fact appear brightly lit if they were seen in contrast to the pitch blackness of real darkness.

If grey stripes are inserted between the white and black stripes, the camera voltage waveform will appear as shown in Fig. 1.3(b). The receiver c.r.t. beam-current during the grey stripes will produce a degree of fluorescence which will appear grey when contrasted against the peak white and black stripes on either side. In monochrome television *all* picture detail is seen as varying levels of illumination between black, through all degrees of grey, up to 'peak white'.

*Synchronisation.* In addition to carrying picture detail, the radiated signal must bear information to keep the receiver scanning spot in step with the camera scanning beam. As in the camera tube, the electron beam in the receiver c.r.t. is deflected by magnetic fields set up in the tube by passing current through deflection coils arranged around the tube neck. The principle is the same as that used in an electric motor; the electron beam behaving as does a current-carrying conductor when situated in a magnetic field. The scanning beam is deflected from side to side by a so-called *line timebase circuit*, and from top to bottom by the *field timebase*.

The spot starts at the top left-hand corner and moves at a uniform speed over to the right, after which it flies back to the left at a much faster speed and proceeds to draw a second line underneath the first. The downward movement of the beam, causing each line to fall beneath the previous one traced, is due to the simultaneous build-up of the vertical deflecting field which continues to increase at a uniform rate until the spot has reached the bottom of the screen. The arrangement of the lines in an interlaced raster has been studied and it is evident that synchronisation information is needed by the receiver at the end of each horizontal line to direct the spot back to the left, and at less frequent intervals a further additional instruction is needed to cause the field timebase circuitry to direct the spot back to the top of the tube again to start its next set of lines.

Fortunately the two distinctly separate kinds of modulation information which the radio wave has to carry, that of picture detail in the one case and synchronisation pulses in the other, are required consecutively and not simultaneously. When the camera and receiver tube beams are moving along a line, voltage variations are required corresponding to changes in brightness due to picture detail along that line. At the end of the line, when the edge of the picture has

been reached, the picture information ceases and it is now a synchronising pulse which is needed to make the receiver spot do what the camera beam now does, i.e. move back to the left of the screen. Again, at the bottom of the screen, when the last line, or half-line of a field has been traced, a pulse suitable for initiating the spot's movement back to the top is needed.

*Flyback.* These speedy return movements of the electron beam to a starting point, following working movements which take place at slower uniform speeds, are referred to as *flyback.* The actual flyback paths traced by the spot on the screen are examined in Chapter 12 and make a most instructive study.

It is now time, however, to look at some block diagrams and to make a broad survey of the whole electronic requirements of a television system.

# 2

# Basic Transmitter and Receiver Arrangements

The transmitter will be outlined only in sufficient detail to establish the way in which the necessary television signal may be built up, but the receiver block diagrams will give a substantially complete picture of the form of a typical modern monochrome receiver. A monochrome receiver is one which is designed to reproduce the normal black and white television pictures as distinct from a colour picture. An understanding of the latter must rest securely upon a thorough familiarity with monochrome principles and circuitry.

**The vision transmitter** (Fig. 2.1)

The way in which the camera produces a line-by-line picture voltage has been described. This voltage is amplified in the 'picture amplifier' and passed on to a combining circuit. A

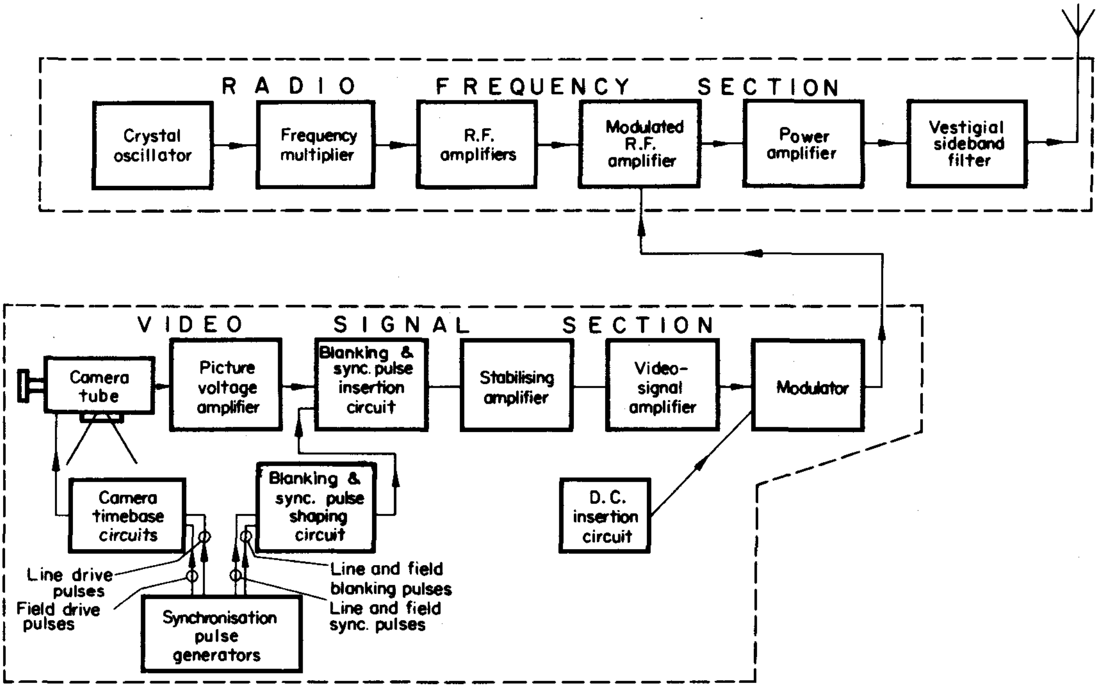



Fig. 2.1 Block diagram showing the basic principle of a vision transmitter

brief summary of the main sections of the transmitter now follows, arranged for convenience in paragraphs bearing headings corresponding to the labels of the boxes shown in the block diagram.

*Camera timebases.* The camera tube deflection coils are fed with appropriately shaped current waveforms from the camera line and field timebases. In both cases it is basically a sawtooth waveform which is needed.

The line timebase must make the spot move at a uniform speed across the screen, and to do this a uniformly increasing magnetic field is needed. During the faster flyback stroke the sole purpose is to get the spot back to its starting point, and this implies that the current must return to its initial value as quickly as possible. A forward to return stroke time ratio of the order of 9 to 1 is usual for this type of timebase circuit, i.e. the flyback occupies about 1/10th of the time duration of one complete scanning cycle. Whilst a constant spot speed during the forward working trace is vitally important, it is clear that this is not so for the return stroke.

In the same way the beam must simultaneously be uniformly deflected downwards by the field timebase, and then returned as quickly as possible to the top of the screen.

*Synchronisation pulse generators.* The timebases must be under the strict control of the synchronising pulse generator circuits. These form the master controlling influence of the entire transmission and reception apparatus. At the end of each line a line synchronising pulse is fed to the line timebase to initiate flyback. A measured interval of time is allowed for the completion of this flyback stroke. Similarly, at the end of each field, a field pulse is needed to initiate the vertical flyback stroke. Again a suitable interval of time is allowed for this movement to be completed. Not only must these generators supply suitably shaped pulses for the direct synchronisation of the camera timebases, but also they must supply the timing pulses from which suitably shaped and timed pulses can be built up to control the receiver timebase circuits.

*Sync. pulse shaping circuit.* Here the required line and field pulse waveforms, which will form the synchronisation information part of the radiated signal, are correctly shaped.

*Sync. pulse insertion circuit.* The complete modulation waveform consisting of the two parts, picture and synchronisation information, is assembled here. Setting up the balance, or comparative levels of these two elements, is an important aspect of the operation carried out in the combining amplifiers at this point. This complete waveform is known as the *video signal*, whilst the end product of the transmitter—a radio wave which is modulated by the video signal—is known as the *vision signal.*

*Stabilising amplifier.* This is a circuit whose task it is to ensure that the video waveform is of standard form. It is capable of accepting and restoring to standard form a video signal which started off as a correctly constituted waveform but became degenerated at some point in the subsequent circuitry.

This must not be taken to imply that the circuit is capable of correcting distortion of picture detail information. Its task is to ensure the correct relative amplitude of the sync. pulses to the picture content and also to reconstitute pulses which have become misshapen.

*Video signal amplifiers.* The video signal is now increased in amplitude from an input level of the order of 1 volt, to the level required for driving the video modulator stage. The latter may require an input of several hundred volts. Wideband amplifiers extending from d.c. to several megacycles are needed. Ideally, to preserve the d.c. component of the video signal, about which more will be said in subsequent chapters, direct coupled amplifiers would be used. However, it is often technically easier and less expensive to use a.c. coupled amplifiers. This involves the loss of the d.c. component which is then re-inserted in a later d.c. restorer or d.c. insertion stage, often called a blanking-level clamp.

*Modulator.** This may be regarded as the output stage of the video signal circuits. It is responsible for feeding the video signal to the modulated R.F. amplifier. A chief requirement of this amplifier is that it should be of low impedance output in order to drive the highly capacitive grid input circuit of the modulated R.F. amplifier without severe attenuation of the higher frequency components of the video signal.

*Crystal oscillator.* A crystal controlled oscillator provides the frequency standard of the transmitter. It is easier to achieve constancy of frequency at a low rather than a very high frequency, and the normal v.h.f. or u.h.f. transmitter technique is to use a crystal working at a sub-multiple of the required frequency. Successive multiplier stages then bring the fundamental frequency provided by the oscillator up to the required level. There is a limit to the multiplying factor possible for any one stage and $\times 3$ and $\times 2$ are common factors employed. A whole series of these may be necessary when the final carrier is at u.h.f.

*R.F. amplifiers.* These increase the amplitude of the R.F. signal to the level required to drive the modulated amplifier stage.

*Modulated R.F. amplifier.* Here the video signal from the modulator stage is made to amplitude modulate the radio frequency input from the R.F. amplifiers. Grid modulation of the modulated R.F. amplifier is often used. This system has a non-linear input/output characteristic. To compensate for the non-linearity it is necessary to stretch the sync. pulses, and logarithmically stretch the picture signal, before the video signal is applied to the modulated amplifier. This pre-distortion is normally carried out in the video amplifier stages.

*Power amplifier.* This stage is responsible for amplifying the amplitude modulated vision signal and developing a radio frequency power output in the aerial system.

*Vestigial sideband filter.* To reduce the frequency space occupied by television signals, part of one sideband of the double sideband A.M. signal produced by the transmitter, is filtered off. This leaves one complete sideband and the 'vestige' of the other; hence the term vestigial sideband transmission. In the case of the 405-line signal, part of the upper sideband is filtered off, but in the 625-line case it is the lower sideband which is so treated. The matter is fully dealt with in Chapter 4.

**The sound transmitter**

This is quite conventional except that the use of the v.h.f. and u.h.f. bands allows for a much wider bandwidth, and hence a much improved degree of fidelity than is possible in the congested bands used for A.M. sound broadcasts. To give the widest possible separation between the sound and vision carriers the sound carrier is always positioned just outside the limit of the fully transmitted vision sideband. With 405-line transmissions the sound carrier is placed 3·5 Mc/s below the vision carrier, but with 625-line transmissions the sound carrier is 6 Mc/s above the vision carrier for the British signal, and 5·5 Mc/s above it in the European signal. A separate sound channel aerial system may be used, but more usually the sound and vision signals are fed to the same aerial via a suitable combining unit.

* The term modulator is used in the field of line transmission and carrier telephony to describe the circuit where the actual process of modulation takes place, i.e. where the modulating information is superimposed upon the carrier.

In radio transmitter technology however, as indicated in the text above, the word modulator is often used to describe the final stage of amplification of the modulating information itself, and the process of modulation is then said to take place in a modulated R.F. amplifier which receives two inputs: the R.F. carrier and the output from the modulator.

It is necessary to be aware of this misleading dual meaning of the word modulator. The line transmission usage appears to be the more logical of the two.

## The television receiver

A detailed examination of 405-line and 625-line waveforms and standards is made in the next two chapters. At this stage, however, it is desirable to make one or two points clear on the question of different line-standards. The first point is that the broad principles of television reception are the same, whatever the standards employed. A technician who is knowledgeable and experienced in one country can be expected to adapt himself with very little trouble to receivers in another part of the world where line standards are different from those to which he has been accustomed.

This work is chiefly concerned with the two systems currently in use in Britain, and for some years to come a close familiarity with both these systems will be necessary to technicians working in this country. In due course 405-line transmissions may disappear from the scene, but in the interim period the added complexity of dual-standards receivers must be faced. On the question of carrier frequencies, at the time when 625-line transmissions were introduced in Britain the v.h.f. Bands I and III were fully taken up by B.B.C. and I.T.A. 405-line channels, and the 625-line transmissions moved into the then unoccupied u.h.f. Bands IV and V. If the 405-line transmissions are finally abandoned, the vacated v.h.f. bands may perhaps be occupied by further 625 transmissions. However, 625-line signals occupy a greater bandwidth than 405-line signals, the corresponding inter-channel spacings allowed being 8 Mc/s and 5 Mc/s respectively, so that fewer 625-line than 405-line channels, can be accommodated in a given band.

The current tendency is to move towards standardisation on 625-lines in many areas of the world and this is of course a desirable development. The basic 625-line system in use is that specified in the C.C.I.R.* standard. Countries adopting this system may introduce minor modifications, as indeed the B.B.C. has done. These will be noted in due course, but they present no difficulty.

Perhaps it is also worth mentioning at this juncture that a colour television signal occupies precisely the same bandwidth as a monochrome transmission of the same number of lines, and the same aerial system handles both with equal efficiency.

Block diagrams of typical television receivers, first a 405-line and then a 625-line, will now be studied. This will give a general idea of the organisation of receiver circuits and will also bring out the main points of difference between the two systems.

## A 405-line receiver (Fig. 2.2)

For convenience the receiver has been divided into main sections which will be dealt with one at a time.

*Aerial system.* The aerials used for television reception are so-called 'resonant' aerials. This implies that they behave in much the same way as tuned circuits, and select one small band of frequencies with a falling response to frequencies on either side. They are clearly not 'high Q' tuned circuits since they are designed to pick up both the vision and sound signals of a given channel. The resonance effect is possible because the carrier frequencies employed are high enough to allow for aerials to be cut to a significant proportion of a wavelength. Usually the active member of the aerial is a 'half-wave' dipole. A check on measurements of practical aerials will reveal that the actual length of this element is a little *less* than half the wavelength of the signal for which it is made. This is because the effective 'electrical length' of a rod aerial is greater than the physical length; the difference getting more as the diameter of rod increases.

* C.C.I.R.: Comité Consultatif International des Radiocommunications.

For average rods used the element is from 4% to 5% shorter than the calculated halfwave dimension.

Basically one aerial is needed for each Band. As far as the 405-line signal is concerned the aim in most parts of Britain is to pick up one 405-line B.B.C. channel on Band I and one 405-line I.T.A. channel on Band III. A separate aerial system is needed for the 625-line signals on Bands IV and V. As a general rule, assuming equal signal strengths, the higher the carrier frequency the greater the number of elements needed to give an adequate signal input to the receiver. This is particularly noticeable in moving to Bands IV and V where multi-element arrays are needed. However, the small dimension of the elements at u.h.f. make this a practical possibility.

To prevent interaction between the Band I and Band III aerials, both of which have to be connected to the same aerial input socket, a filter unit known as a *diplexer* is used. This has the property of allowing both the Band I and Band III signals through to the common outlet cable but effectively blocks signals from either aerial from reaching the other. It consists of a low pass filter between the Band I aerial and the common outlet socket and a high pass filter between the Band III aerial and this socket. This is placed as near to the aerial installation as possible and a single coaxial cable down-lead completes the connection to the receiver. Many dual-band aerials are made for use in areas where the signal strength is adequate. This is cheaper than two separate aerials and obviates the necessity for a diplexer.

*The tuner unit.** The purpose of the tuner unit is to amplify both the sound and the vision signals picked up by the aerial and to convert the carrier frequencies into the chosen intermediate frequencies in use in the receiver.

The first stage is an R.F. amplifier of bandwidth sufficient to accept both the vision and the sound signals at one and the same time.

The second stage is the mixer which receives three inputs; the amplified sound and vision signals from the R.F. amplifier, and the output from a local oscillator. The latter beats in the mixer with both the carrier frequencies simultaneously to produce two intermediate frequencies; the sound I.F. and the vision I.F. These correspond to the difference between the oscillator frequency and the sound and vision carrier frequencies respectively. The standard intermediate frequencies in use in this country in 405-line receivers are:

| | |
|---|---|
| Vision I.F. | 34·65 Mc/s |
| Sound I.F. | 38·15 Mc/s |

If a tuner unit is switched to Channel I in a receiver using these I.F.'s, the local oscillator will be working at 79·65 Mc/s. The carrier frequencies on this channel are 45 Mc/s for Vision and 41·5 Mc/s for Sound. The required I.F.'s are then produced as follows:

(Oscillator frequency of 79·65 Mc/s) − (vision carrier frequency of 45 Mc/s) = Vision I.F. of 34·65 Mc/s

(Oscillator frequency of 79·65 Mc/s) − (sound carrier frequency of 41·5 Mc/s) = Sound I.F. of 38·15 Mc/s

Notice that the frequency changing process reverses the relative positions of the sound and vision signals. This is obvious since the oscillator works above the signal frequencies, and the

* A knowledge of superhet principles has been assumed by the author. Those whose knowledge is limited, or who are perhaps 'rusty' are advised to combine some general revision studies with their television reading. The specialised v.h.f. and u.h.f. techniques employed in television tuner units will of course be dealt with subsequently.

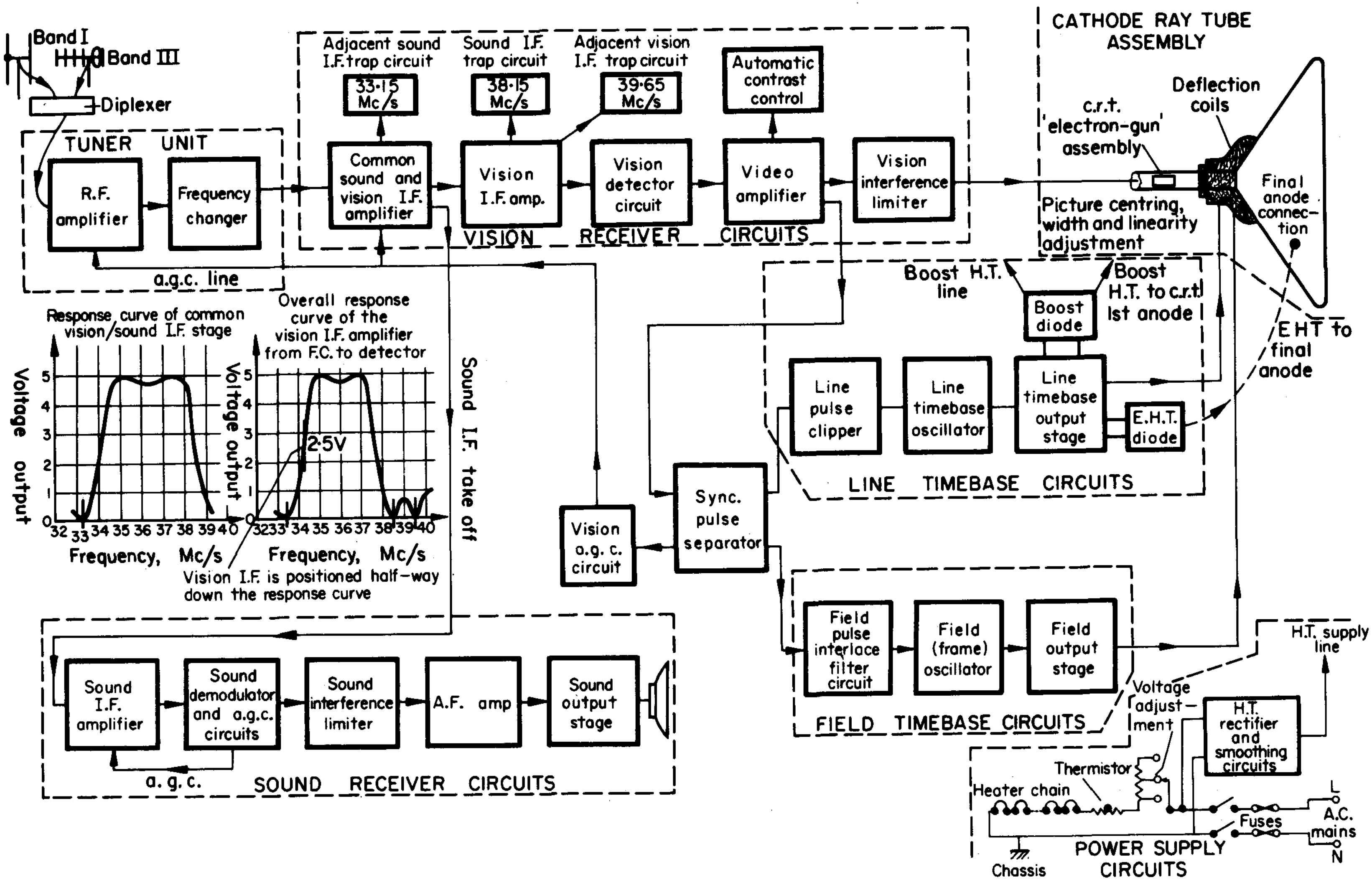



Fig. 2.2 Block diagram of a 405-line television receiver

'difference' frequencies produced when the sound and vision carrier frequencies are subtracted, must give a higher I.F. from the lower frequency sound signal.

It is worth pausing to note that the vision and sound signals would only remain in the same relative position, that is with the vision carrier frequency higher than the sound carrier frequency, if the oscillator frequency were pitched *below*, instead of *above* the carriers. The need for this did arise in the early days after the introduction of Band III programmes in Britain. In order to enable single-channel Band I receivers, some of which were 'straight sets', to receive the second Band III programme, converters were manufactured which translated the required Band III transmission to the Band I frequencies corresponding to the receiver's single channel. The converter 'Band I/Band III' switch simply switched the receiver's aerial input socket either direct to the Band I aerial or to the output of the converter. A separate Band III aerial was connected to the converter input circuit. The converter consisted of an R.F. amplifier and a frequency changer, and it was muted on Band I by having its H.T. supply switched off.

As an example, to illustrate this channel-frequency translation process, suppose the converter were designed to translate Band III, Channel 9, down to Band I, Channel 1.

The frequencies involved are:

| *Band III, Channel 9* | | *Band I, Channel 1* | |
|---|---|---|---|
| Vision carrier | 194·75 Mc/s | Vision carrier | 45 Mc/s |
| Sound carrier | 191·25 Mc/s | Sound carrier | 41·5 Mc/s |

The oscillator frequency must be placed so that it is 45 Mc/s below the Channel 9 vision frequency, which of course automatically places it 41·5 Mc/s below the Channel 9 sound frequency. Hence:

| | | | | |
|---|---|---|---|---|
| Band III vision frequency of 194·75 Mc/s | − | oscillator frequency of 149·75 Mc/s | = | Band I vision frequency of 45 Mc/s |
| Band III sound frequency of 191·25 Mc/s | − | oscillator frequency of 149·75 Mc/s | = | Band I sound frequency of 41·5 Mc/s |

Continuing with the Tuner Unit, the anode, or output circuit, of the mixer contains a tuned circuit of bandwidth wide enough to accept both the vision and sound I.F.'s simultaneously, and these signals are passed on to the next stage.

The tuning of the R.F. and oscillator tuned circuits is pre-set, and the tuner unit is switched from channel to channel. Modern tuner units are remarkably stable in performance and usually it is only the oscillator pre-set tuning which needs adjustment from time to time; for example when valves are changed. A user's 'fine-tuner' control provides a small adjustment of oscillator frequency in most receivers, but some receivers on the market are not provided with a manual control of oscillator frequency at all. There is no doubt of the desirability of putting control of oscillator frequency out of reach of the user, provided the oscillator may be relied upon to remain completely stable once having been correctly set by the technician. This will be perceived more clearly when the technical effect of mistuning the oscillator is studied in subsequent work. Whether or not present day tuner unit performance justifies the omission of the fine-tuner is still a matter of debate.

*Vision receiver circuits*

The purpose of this section is to amplify the vision I.F., demodulate it, and amplify the recovered *video* signal for onward transmission to the cathode ray tube on the one hand, and the timebases on the other.

Often the first I.F. amplifier is used as a common sound and vision stage, with a broad response curve capable of accepting both these signals at the same time. The sound I.F. is then extracted from the output from this common I.F. amplifier by inclusion of a suitable 38·15 Mc/s tuned circuit in its anode inter-stage coupling circuitry, and passed on to the sound receiver. Alternatively the 'sound I.F. take-off circuit' directly follows the tuner unit and no common I.F. amplifier is used.

The block diagram shows a sketch of the broad response curve of the common stage and it will be seen that both the 34·65 Mc/s vision I.F. signal and the 38·15 Mc/s sound I.F. signal are accommodated by it.

In the I.F. amplifier circuitry, provision must be made for the rejection of signals from adjacent channels, and for this purpose special tuned circuits, called trap-circuits, are connected in the signal path in such a way that the offending frequencies are removed. These trap-circuits are disposed at convenient places in the I.F. amplifiers. Their position will vary from receiver to receiver, but the first one, which may be either in the output circuit of the first I.F. amplifier, or immediately following the tuner unit mixer, is usually the 'adjacent-sound I.F.' trap circuit tuned to 33·15 Mc/s.

The way in which these unwanted adjacent channel I.F. signals appear is illustrated in Fig. 2.3. As an example, suppose that the receiver is switched to Channel 3 on Band I. The

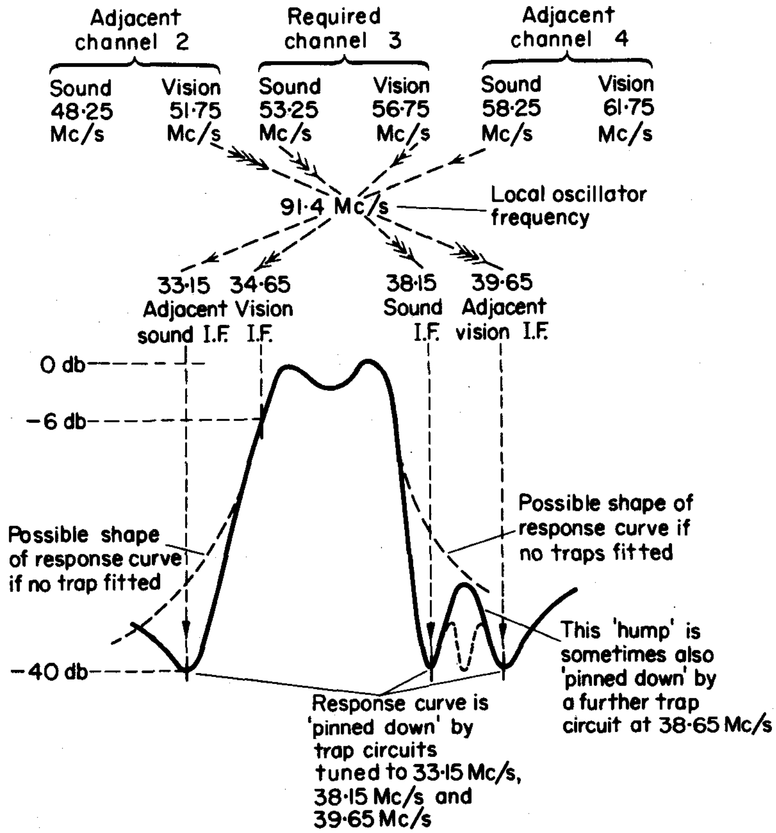



Fig. 2.3 Overall vision I.F. response curve for a 405-line receiver assumed tuned to Channel 3
The diagram shows the disposition of intermediate frequencies arising from this and the adjacent channels.

oscillator frequency is found by adding 34·65 Mc/s to the vision carrier, which for Channel 3 is 56·75 Mc/s, giving an oscillator frequency of (56·75 + 34·65) Mc/s = 91·4 Mc/s.

In the diagram the I.F.'s which result from the wanted Channel 3 and the unwanted 'Adjacent Channels' 2 and 4 are found by following the dotted lines from the signal frequencies through the oscillator frequency to the appropriate I.F.; the latter being the difference between the oscillator and signal frequencies in each case. This diagram brings out clearly the way in which sound and vision carriers reverse their respective positions after passing through the frequency changing process. The trap circuits, which are seen effectively to 'pin down' the I.F. response curve, play an important part in the shaping of the final I.F. response curve to its necessary form.

Examination of the response curve sketched in Fig. 2.3 and on the block diagram, Fig. 2.2, will show that the vision I.F. of 34·65 Mc/s is placed some distance down the low frequency skirt of the response curve. Its actual position is half-way down if the '$y$-axis' is scaled in volts (or voltage gain) as it is in the sketch on the block diagram; but 6 dB down if the axis is scaled in decibels. The reason for this positioning arises from the use of a vestigial sideband signal, and the matter is explained in Chapter 4.

Following the common I.F. stage is a further stage (or sometimes two stages) of vision I.F. amplification. The use of high gain frame-grid valves for R.F. and I.F. amplification has made it possible to achieve adequate gain with just the common stage and one further stage of I.F. amplification. Earlier receivers often employed an additional I.F. stage making three between the tuner unit and the vision detector. This may still be found in some very sensitive 'fringe' receivers which have to work on very low input signal levels and hence need the extra gain.

Included in this circuitry must be a sound I.F. trap circuit tuned to 38·15 Mc/s to eradicate this signal from the vision I.F. path. If this is not removed, the sound I.F. signal will be demodulated by the vision detector, and the A.F. modulation passed on via the video amplifier stage to the cathode ray tube giving rise to the well known phenomenon of 'sound-on-vision'.

Following the last vision I.F. amplifier is the vision detector circuit which demodulates the vision signal and recovers the video signal. At this point the latter will have an amplitude of the order of 1·5 to 5 volts. The circuit must be carefully positioned to minimise radiation to earlier stages in the receiver of unwanted frequency components produced by the demodulation process. Often the detector is completely screened for this purpose. The vision signal is amplitude modulated and the detector follows normal A.M. detector principles, but the component values and circuit arrangements differ from those found in normal broadcast receiver practice because of the much higher I.F. employed in television, and the very much higher frequency components present in the video modulation.

The video amplifier increases the amplitude of the video signal to the order of 50 to 75 volts. It is a wideband amplifier capable of handling a frequency range which, for the 405-line signal, must extend from d.c. to 3 Mc/s. The justification for the choice of 3 Mc/s as the upper frequency aimed at in the case of this particular line standard is dealt with in Chapter 3. The response is often made to dip sharply to zero at 3·5 Mc/s to remove an 'inter-I.F.' beat-frequency component which is almost inevitably present.

Between the video amplifier and the cathode ray tube is an interference limiter circuit which has the task of preventing the sudden voltage surges, caused by interference sources such as vehicle ignition, from reaching the cathode ray tube where they would give rise to random white blobs across the picture. In practice the circuit cannot entirely remove these without seriously degrading the picture, but they can be reduced to small sharp spots or dashes of

light, or as is sometimes preferred, inverted by so-called 'black spotter circuits' to form black interference marks instead of white ones.

An automatic contrast control circuit is sometimes included between the video amplifier and the cathode ray tube. The purpose of this is to give an automatic adjustment of the general brightness of the picture, to compensate for changes in the level of ambient room lighting.

*Sync. separator circuit.* This stage also receives the video signal from the video amplifier and has the job of stripping off the synchronisation pulses from the composite waveform. These, duly processed, are fed off to the two timebase circuits.

*Line timebase.* The line pulses are suitably sharpened and shaped by a line pulse clipper or similar circuit, and passed on to control the frequency of the line timebase oscillator. Manual control of the line oscillator frequency by means of the 'line-hold' control enables the user to set the frequency to a point at which the arriving sync. pulses are able to get a grip on the oscillator and lock it to the correct frequency.

The line output stage has the primary task of setting up the correct sawtooth scanning current in the deflector coils. It is the most hard-worked stage in a receiver, and much careful development work has been carried out to produce the highly efficient circuits necessary to cope with the demands made upon this stage.

Included, as integral parts rather than adjuncts of this stage, are the boost diode circuit, and the E.H.T. circuit. The former provides the line output stage with an H.T. supply considerably in excess of that existing on the main H.T. line. The circuit actually adds a voltage of from 300 to 500 volts to the normal H.T., to give a boost H.T. of from 400 to 700 volts. This voltage is also convenient for a cathode ray tube 1st anode and focus anode H.T. supply and is utilised in other parts of the receiver where an added voltage is advantageous. The E.H.T. circuit provides the 'extra-high-tension' voltage of perhaps 15 to 17 kilovolts, which is needed for the final anode of the tube.

### *Field (or 'frame') timebase circuit*

The field synchronisation pulses must be identical on odd and even fields if good interlace is to be achieved. An interlace filter circuit processes the field pulses arriving from the sync. separator, to ensure that this is so.

The field oscillator frequency is set by the user to achieve vertical lock by adjustment of the appropriate hold control, which despite what has been said on the subject of the word 'frame' may sometimes continue to be called a 'frame hold' rather than a 'field' or 'vertical' hold control for some years to come. An oscillator output amplitude control allows for adjustment of picture height.

The field output stage is a conventional power amplifier stage with a less arduous task to perform than the line output stage. Its task is to set up the correct 50 c/s sawtooth current waveform in the vertical deflection coils. Any complexity found in this circuit is associated with linearisation of the raster.

### *Cathode ray tube assembly*

The deflection coil assembly mounted on the neck of the cathode ray tube often includes arrangements for adjustment of picture width and line linearity, as well as for picture centring.

Small magnets are usually found which serve to 'pull-out' the sides of the raster which otherwise, with modern wide-angle tubes, tends to have concave sides to give what is often described as a pin-cushion shape. The arrangement sounds, and is, rather arbitrary but the end product is a satisfactory raster which after all is what matters.

Also provided is a facility for focusing the spot and this may involve adjustment of the H.T. potential to the focus anode in the case of current electrostatically focused tubes, or adjustment of a ring magnet (usually a 'permanent' magnet but sometimes 'electromagnetic') in the case of older magnetically focused tubes.

*Vision a.g.c.*

Automatic gain control of the vision receiver is essential to minimise the necessity for the user to be constantly adjusting the panel contrast control to compensate for the changes in level of input signal-strengths which are always taking place; not only when switching from one channel to another but also due to constantly changing propagation conditions.

The circuits developed for this purpose are many and various and examples of these will be studied in due course.

In the block diagram the a.g.c. circuit is shown as stemming from the input circuitry of the sync. separator. This is often the case but it need not be so as will be seen when a.g.c. circuits are examined.

*Sound receiver*

This is very much a conventional sound receiver superhet circuit. The major difference between it and a normal A.M. broadcast receiver lies in the very much higher I.F. which, as has been stated, is 38·15 Mc/s instead of the normal I.F. of 465 kc/s.

The bandwidth of the television sound receiver I.F. stage(s) is also very much wider, being in the region of 200 kc/s rather than of the order of 10 kc/s. Clearly this is a much greater bandwidth than is needed to accommodate the sidebands of a signal which is modulated only by audio frequencies, but there are good reasons for providing this broad response. By so doing the effect of tuner unit oscillator drift does not result in distortion as it would do if the response curve were tight, since in this case the I.F. set up by the received sound carrier could move down the skirt of the I.F. response curve, or even completely outside it.

A second reason for the broad response is connected with rejection of sound interference, such as ignition noise. Such interference, which is known as impulsive noise, causes high-amplitude short duration pulses, or 'spikes', to be superimposed on the modulation envelope. Paradoxically, the steep-fronted nature of these pulses must be preserved in the I.F. amplifier, in order to allow for their easy removal from the A.F. waveform after the detection process. This is because a sharp pulse contains a wide range of frequency components and a broad-bandwidth is needed to pass these. A narrow passband I.F. amplifier would cause the pulses to become broad based and stubby in appearance. Removal of these without severe distortion of the A.F. signal they were 'riding on' would be impossible. On the other hand, if the pulses retain their original sharp spike-like form, it is comparatively easy to devise circuitry which can distinguish between these and the A.F. signal, and therefore remove them efficiently, without seriously impairing the quality of the sound.*

Because of the broad bandwidth which is available the degree of fidelity of television sound should be much better than normal A.M. broadcasts and since this is so it seems a very great pity that tiny inadequate loudspeakers are all too often fitted in domestic receivers.

The sound I.F. is demodulated in a normal A.M. detector circuit. Ignition and other staccato

* It is also interesting to note that, even if a narrower bandwidth, e.g. 20 kc/s, were desired, the 38·15 Mc/s tuned circuits used in the receiver sound I.F. stages would require Q-factors so high, e.g. of the order of 2000, that crystal filters would be needed! Normal tuned circuits having Q-factors of the order of 100 to 200, however, yield bandwidths in the region of 200 kc/s or more with an I.F. of 38·15 Mc/s.

interference noises are then limited by a sound interference limiter, placed between the detector and the A.F. amplifier stage. The A.F. amplifier and output stages are conventional.

*Power supplies*

Power supply arrangements are usually quite straightforward. To avoid the use of heavy and costly mains transformers, the majority of television receivers employ 'transformerless' techniques. The valve heaters are arranged in a series chain with a suitable tapped voltage dropping resistor. A series thermistor is fitted to protect the heaters during the warming-up process, but in this direction valve manufacturing techniques of recent years have produced so-called 'equalised heaters' which lessen the necessity for this protective component.

**A 625-line receiver** (Fig. 2.4)

Much of what has been said about the 405-line receiver is equally applicable here. A study of the block diagram will reveal a close similarity between the two receivers. In the text which follows, attention is directed at the main points of difference.

*The tuner unit.* The principle of this, and the function it performs is exactly the same as before. The difference as far as British receivers are concerned is that, at the time of writing, the incoming carrier frequencies are all in the u.h.f. Bands IV and V. At this point it may be instructive to specify the frequency ranges of the various Bands; the figures quoted being those used in Europe. The limits of the Bands are slightly different in the U.S.A. and slightly different again in the Far East and Australia.

| | | | |
|---|---|---|---|
| Band I | (v.h.f.) | 41 to 68 Mc/s | (contains British television channels 1 to 5) |
| Band II | (v.h.f.) | 87·5 to 100 Mc/s | (used in Britain for F.M. sound broadcasting) |
| Band III | (v.h.f.) | 174 to 216 Mc/s | (contains British television channels 6 to 13) |
| Band IV | (u.h.f.) | 470 to 610 Mc/s | ( „ „ „ „ 21 to 34) |
| Band V | (u.h.f.) | 610 to 940 Mc/s | ( „ „ „ „ 39 to 68) |

The circuit principles employed in u.h.f. R.F. amplifiers, mixers and oscillators are similar to those in v.h.f. tuners, but specialised circuit techniques are necessary when working at frequencies of this high order. The familiar tuned circuit consisting of an inductor and capacitor, already barely recognisable in Band III v.h.f. circuitry, now disappears in favour of resonant transmission line sections or 'Lecher' wires.

As stated earlier, with 625-line transmissions the sound carrier is positioned above the vision carrier, instead of below it as in the 405-line system.

The vision I.F. employed is 39·5 Mc/s, and with the oscillator working above the carrier frequencies, the usual reversal of positions takes place and the sound signal leaves the mixer as an I.F. which is 6 Mc/s lower than the vision I.F., i.e. at 33·5 Mc/s.

As an example, Fig. 2.5 shows the frequency-changing process when a tuner unit is set to receive Channel 44 on Band V. The diagram also shows what adjacent channel I.F.'s are produced with the British 625-line system.

*The vision and sound receiver circuits*

The vision I.F. of 39·5 Mc/s together with the sound signal at 33·5 Mc/s are fed to the first vision I.F. amplifier. There may be either two or three stages of vision I.F. amplification, and a sketch on the block diagram shows the form of the overall vision I.F. response curve. The '$y$-axis' has again been scaled in voltage output on this sketch, but in Fig. 2.5, which also shows the same response curve, the scale is in decibels with the top of the curve taken as a

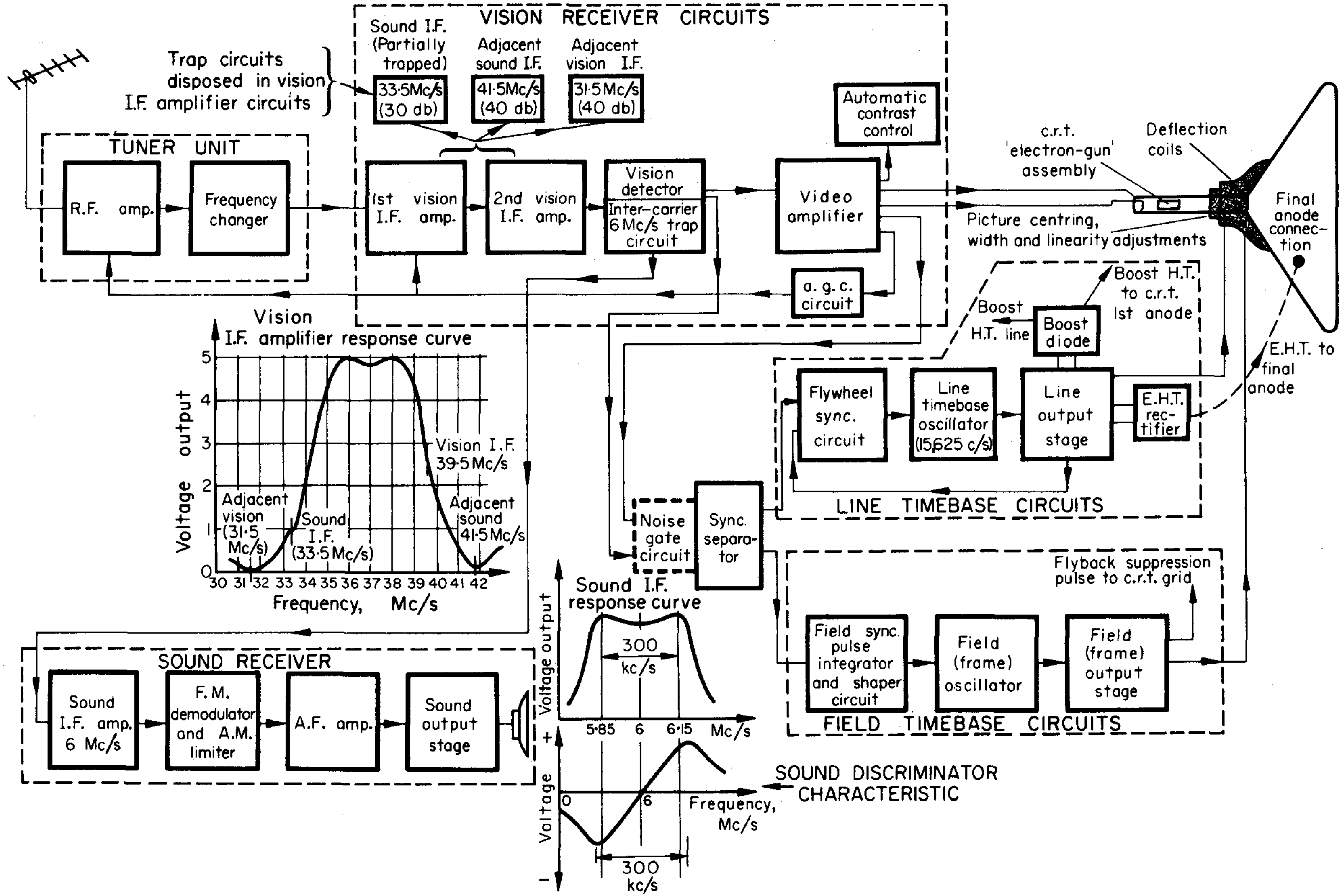



Fig. 2.4 Block diagram of a 625-line television receiver

reference level of 0 dB. The use of the logarithmic units 'decibels' instead of simple voltages (or voltage gains) makes it possible to accommodate a much wider range of levels on the $y$-axis.*

It will be observed that the dB scale has the advantage of allowing attention to be drawn to the desirable 'relative attenuation' of the various frequencies encountered by the I.F. amplifier.

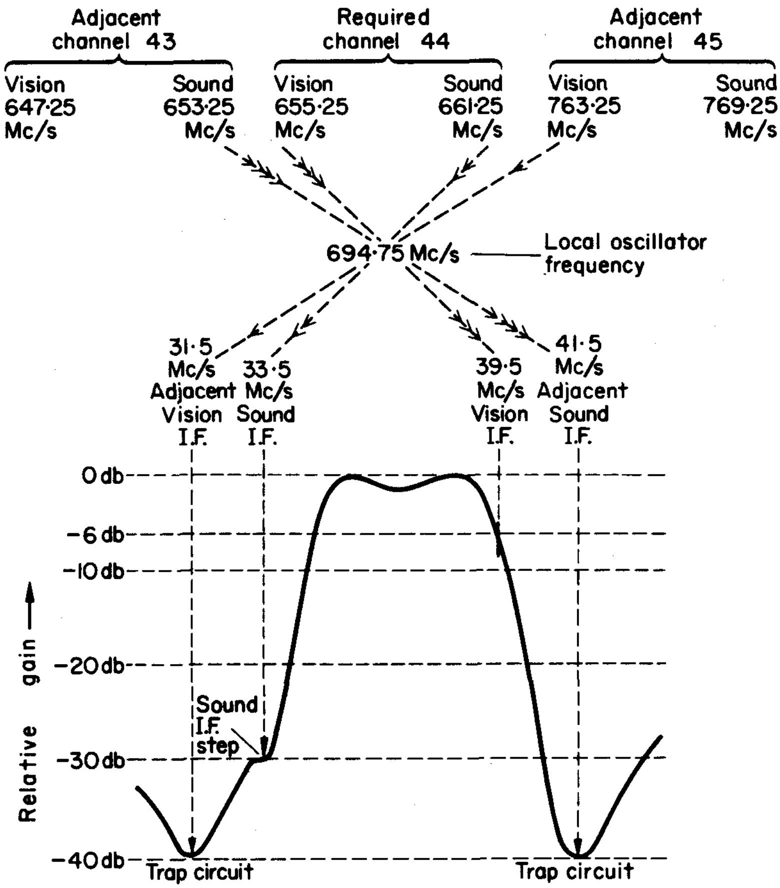



Fig. 2.5 Overall vision I.F. response curve for a 625-line receiver assumed tuned to Channel 44 on Band V
The diagram shows the disposition of intermediate frequencies arising from this and the adjacent channels.

* For those not very familiar with decibels, it should be noted that if the maximum voltage output (or reference level output) is specified as 0 dB, then the level of $-6$ dB corresponds to an output voltage of only one-half the maximum level. Similarly $-40$ dB corresponds to only 1/100th of the maximum level.

In general, for the purposes of the curves shown, the relative voltage output (or relative voltage gain) at a given frequency $f'$, may be computed by using the formula:

$$\text{Relative voltage output (or gain)} = -\left\{20 \log_{10} \frac{\text{Maximum output (or gain)}}{\text{Output (or gain) at } f'}\right\} \text{dB}$$

It is immaterial to the calculation, whether absolute voltage outputs or numerical voltage gains are stated, provided of course that either the one or the other is used in both numerator and denominator.

The first point to notice is that the comparative positions of all sound and vision frequencies are reversed when compared with the 405-line signal. This is obviously the case since the sound is transmitted on opposite sides of the vision signal in the two systems. The vision I.F. is positioned 6 dB down the high frequency side of the I.F. response curve in the 625-line receiver, whereas it is 6 dB down the low frequency side in the 405-line receiver. Similarly, the interfering frequency on the low frequency side is now an 'Adjacent Vision' I.F. instead of 'Adjacent Sound' I.F. and so on.

The second point to notice is the position of the sound I.F. under this response curve. It is shown as being 30 dB down the low frequency side of the curve on Fig. 2.5, and the position is characterised in typical receiver response curves by what is sometimes referred to as the sound I.F. 'step'. This draws attention to an important difference between the 625-line and 405-line systems.

Unlike the 405-line sound transmission which is Amplitude Modulated, the 625-line sound carrier employs Frequency Modulation. This gives rise to a rather different arrangement in the way in which the sound signal is handled in the 625-line vision I.F. amplifiers.

In the 405-line receiver the sound I.F. is extracted, as was explained, either immediately following the tuner unit or after the first 'common' I.F. amplifier stage. Thereafter it is necessary to ensure, by means of sound I.F. trap circuits, that no remaining sound I.F. signal travels through the rest of the vision I.F. amplification stages to reach the vision detector. This precaution is necessary because both the vision and sound signals are amplitude modulated, and naturally the vision A.M. detector will demodulate both signals with equal dexterity and pass the modulation on through the video amplifier to the c.r.t.

However, in the 625-line system, no such problem arises because the vision carrier is an A.M. signal and the sound is F.M. The vision detector will not demodulate the sound signal, provided it is not 'swamped' by it when of course peculiar effects can occur. This is fortunate because it enables another problem to be overcome.

In due course sound F.M. demodulation circuits will be studied. At this stage it is sufficient to state that it is extremely important, if distortion is to be avoided, that the Intermediate Frequency presented to the F.M. demodulator remains constant at the frequency to which the latter is tuned. If the demodulator were tuned to the sound I.F. of 33·5 Mc/s, trouble would be experienced because of tuner unit oscillator drift, which could quite easily take the I.F. far enough away from its correct value to cause severe sound distortion. It should be noted that picture quality is not noticeably affected by drifting of the order discussed here.

*Inter-carrier sound.* The solution is rather clever. Instead of tuning the sound receiver demodulator to the sound I.F. produced by the mixer, it is tuned to the 'Inter-carrier beat frequency' of 6 Mc/s. Sufficient sound I.F. is then allowed through the vision I.F. amplifier to give rise to a second frequency changing effect at the vision detector. Any non-linear device will produce sum and difference frequencies when presented with voltages at two different frequencies. For example, u.h.f. and micro-wave 'mixers' are simply crystal diodes, which when fed with signal and oscillator voltages, produce the required I.F. currents. The vision detector diode treats the two I.F.'s reaching it in exactly the same way. The difference frequency *must be* 6 Mc/s, since whatever the oscillator does, and however much the vision and sound I.F.'s depart from their normal value because of it, the incoming vision and sound carriers remain correctly spaced and the two intermediate frequencies are *always* 6 Mc/s apart.

Like any other 'intermediate frequency' the 6 Mc/s signal carries the modulation borne by the frequencies giving rise to it. It therefore carries the F.M. from the sound I.F., and of course some A.M. information from the vision carrier. This 6 Mc/s signal is now treated as the main

sound receiver I.F. and passed to a 6 Mc/s I.F. amplifier. This has a bandwidth of 300 kc/s which is adequate to handle the bandwidth of the F.M. signal.

*F.M. demodulator.* Following the 6 Mc/s sound I.F. amplifier is the F.M. demodulator circuit. There are several possible types of circuit available for this task but all have the basic frequency discriminator characteristic sketched on the block diagram. This shows that if the 6 Mc/s I.F. is swinging to and fro in frequency about its centre value of 6 Mc/s (i.e. is frequency modulated), the discriminator will give rise to an output voltage which alternates in value above and below the zero voltage point at the same frequency as that at which the carrier swings to and fro. This frequency is of course the audio frequency so that the discriminator output is an A.F. voltage. The louder the original sound, the wider does the 6 Mc/s I.F. swing either side of its centre value, and the greater is the amplitude of the output voltage.

*A.M. limiting.* The 6 Mc/s I.F. carrier also inevitably carries A.M. variations; both in the form of video information from the vision I.F. and interference from ignition and other sources. The sound receiver must include provision for stripping off these amplitude variations. Circuits for this purpose are called limiters. Some forms of F.M. demodulators also function simultaneously as amplitude limiters, and these forms naturally find favour because they avoid the use of a separate limiter stage. The block diagram indicates the presence of a dual purpose circuit.

*Positive and Negative modulation.* In the next chapter video waveforms are studied. At this stage, however, it is convenient to introduce a further difference between the 405 and 625-line systems.

In the former signal, increases of picture brightness cause increases of carrier amplitude. Such a signal is said to be positively modulated. The maximum carrier amplitude, referred to as 100% modulation, corresponds to maximum brightness, and all other lower degrees of brightness produce successively smaller carrier amplitudes, down to the level reached for black parts of the picture. This 'black level' carrier amplitude is actually 35% of the maximum peak white level referred to; not zero as perhaps at first may be imagined. The reason for this hinges upon the necessity to so arrange the two aspects of the video modulation, that is to say the picture and synchronisation components, that they may easily be separated at the receiver. To facilitate this, all synchronisation information is contained between carrier amplitude levels of 0% to 30% whilst the picture detail lies between the 35% and 100% levels. The sync. pulses take the carrier down to zero level.

With the 625-line signal the position is reversed and synchronisation pulses take the carrier up to 100% modulation and are contained within the limits 77% to 100%. From the details given for the 405-line signal it will be seen that the borderline between synchronisation pulse information and the picture detail is the region of black level. Moving further away from this black level on the opposite side to that occupied by the sync. pulses, corresponds to increasing picture brightness. The same thing applies with the 625-line signal, but here the sync. pulse tips are at 100%, black level is at 77%, and increasing brightness drives the carrier amplitude *down* from 77% towards 0%. For reasons to be discussed, maximum brightness, or 'peak white' as it is usually called, is fixed at 18% carrier amplitude and not 0%. A signal in which increasing brightness causes decreasing carrier amplitude is said to be negatively modulated.

To illustrate this, Fig. 2.6 shows sketches of positively and negatively modulated vision signals.

Naturally this difference in direction of modulation causes some differences in the circuitry for processing the video signal. However, these are not very great, and to demonstrate this the video signals modulating the carriers in Fig. 2.6 are assumed identical. The only difference between the two is seen to be simply that the video waveforms impressed upon the carriers are 'opposite ways up'. *Either* a positive-GOING *or* a negative-GOING video waveform may be

obtained from *each* of the signals, simply by reversing the diode in the simple detector circuit shown. It must be stressed that in this context, the expression positive-going signal implies that increasing levels of brightness drives the video signal voltage in a positive direction relative to its value at black level. Whilst this is true of both the waveforms labelled positive-going in the

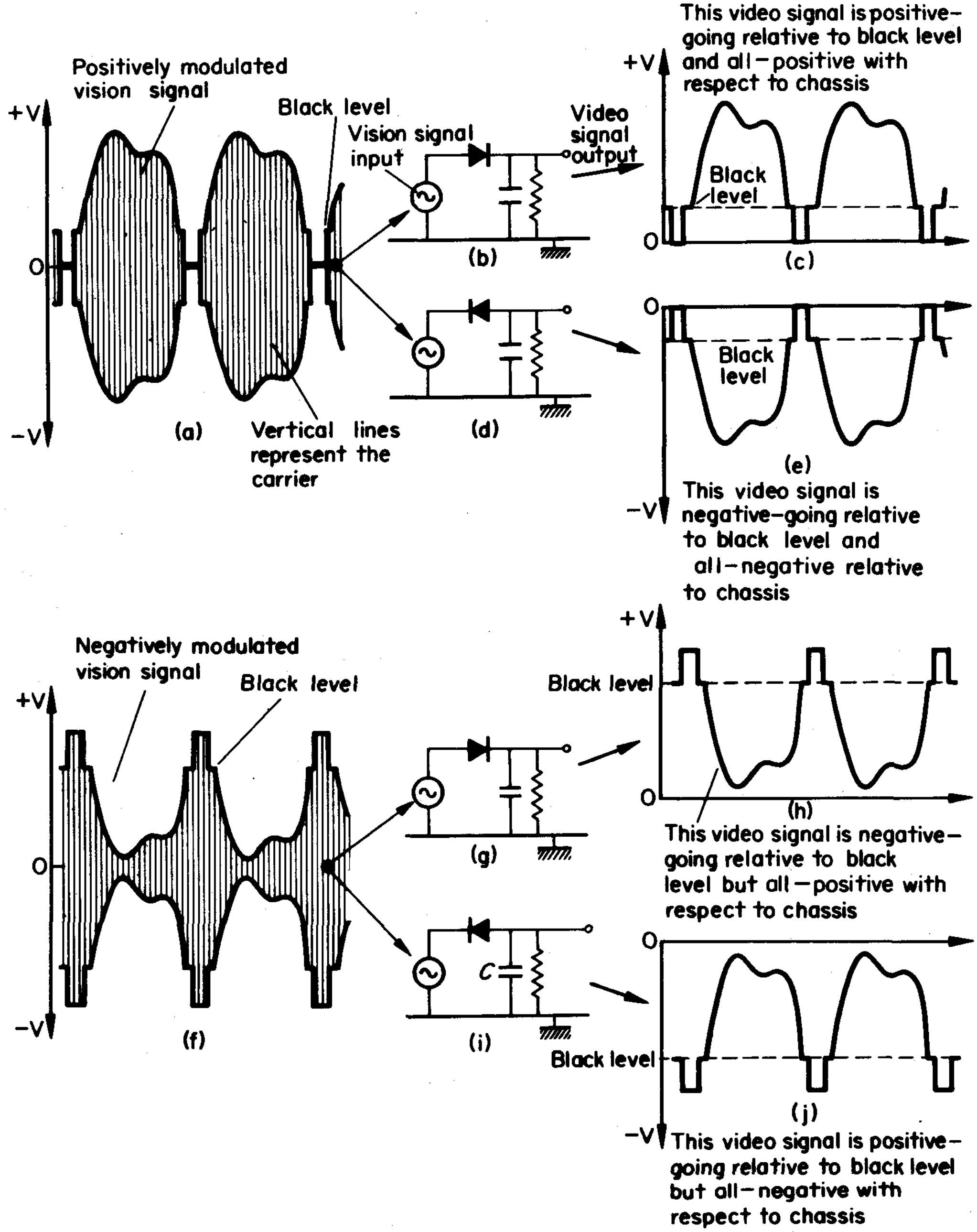



Fig. 2.6 These diagrams show the polarities of the video signals which are produced when positively and negatively modulated vision signals are applied to a simple detector circuit

diagram, it will be noticed that in one case the *whole of the waveform* is positive with respect to 'chassis' but in the other the whole of it is negative to chassis. (See (c) and (j).)

If the polarity of the waveforms were described *relative to chassis*, one would be said to be 'all-positive' with respect to chassis and the other 'all-negative'. They are none the less *both* positive-going in terms of picture voltage *relative to black level.* A voltage which moves from say $-10$ V to $-6$ V is positive-going even though both voltages are negative.

This matter need not be unduly worried about at this juncture but the fact that the 625-line video waveform is below chassis potential, whilst the 405-line waveform is above it, does give rise to necessary differences between 625- and 405-line receivers in both the detector-to-video amplifier coupling methods and in the video amplifier bias arrangements. This question is fully discussed in Chapter 5.

One of the main technical differences between positively and negatively modulated signals centres upon the effects of amplitude interference such as ignition noise. Such interference causes erratic and sometimes devastating changes in the maximum carrier amplitude. With positive-modulation, a steep increase in carrier amplitude drives the carrier beyond peak white, with the resulting white blobs on the screen which were spoken about earlier. In the case of the negatively-modulated signal it is the sync. pulses which exist at maximum carrier amplitude, and the effect of interference is to both mutilate some of these, and to produce a whole lot of spurious random pulses. This, unless something is done about it, completely upsets the synchronisation of the receiver timebases.

### *Synchronisation and timebases*

As a first measure to obviate this effect a 'noise gate circuit' may be inserted as an adjunct to the normal sync. separator. The purpose of this is to 'switch off' the sync. separator when interference pulses drive the carrier amplitude above the normal 100% level. In the block diagram of Fig. 2.4 the dotted line from the vision detector indicates that the noise gate circuit is in this case activated by noise pulses derived from this source.

It is worth mentioning that if transmissions are in the u.h.f. region, such interference is very much less than at lower frequencies. There is a very marked lessening of interference even when moving from Band I to Band III in the v.h.f. ranges, and the step-up to u.h.f. provides a welcome relief from all but the worst of this trouble. When u.h.f. is used for 625-line working, therefore, and the signal strength is adequate, it is possible to do without the noise gate circuit.

This is in part due to the second step which is taken to protect the receiver from random sync. pulses. This involves the use of 'flywheel synchronisation' of the line timebase instead of 'direct' synchronisation. In the latter technically easier and cheaper system, the line timebase oscillator is triggered line by line by the incoming line sync. pulses. The absence of pulses or the existence of an extra random pulse, both of which effects can result from interference, is bound therefore to cause a momentary loss of synchronisation which gives rise to the symptom known as 'line tearing'.

With flywheel synchronisation the line timebase oscillator is put under the control of a local circuit whose controlling influence depends upon the arrival of a succession of pulses instead of upon *each* single one. It is not affected by the appearance of a random pulse, or the absence of a pulse but only by a sustained alteration to the regular supply of pulses which activates it. The circuit may be visualised as having 'electrical momentum' which so conditions it that it will take no notice of transitory changes.

The block diagram indicates that the flywheel synchronisation circuit receives two inputs; one from the sync. separator and the other in the form of a reference waveform derived from

the line output stage. The purpose of this, and the mode of operation of various flywheel synchronisation arrangements, will be studied subsequently.

It should be mentioned that many 405-line receivers also employ flywheel circuits to enhance their performance in fringe areas where the 'signal-to-noise' ratio is poor. To sum up in general terms, it may be said that flywheel synchronisation is sometimes a desirable additional facility on receivers handling positively modulated vision signals, but is usually a vital necessity for receivers working on a negatively modulated system.

As indicated in Chapter 1, the line oscillator frequency for the 25 pictures-per-second 625-lines-per-picture system is $625 \times 25 = 15{,}625$ c/s. The line output stage performs in the same way as for the 405-line system with modifications occasioned by the higher line frequency.

*Interlacing.* The 625-line waveform contains a sequence of pulses before and after field sync. pulses, which are not present in the 405-line waveform. These are called 'equalisation pulses' and their presence makes it somewhat easier to achieve good interlacing without the inclusion of special interlace filter circuitry. Accordingly the 405-line block diagram shows the presence of an interlace filter circuit, but this has been omitted on the 625-line diagram and replaced by a pulse-shaping circuit. These matters are investigated further in later chapters.

The 50 c/s field oscillator and output stage are of course doing the same task as in the 405-line receiver.

*Dual standards receivers*

Typical 405-line and 625-line receiver arrangements have been studied separately. As a complication, all the while both systems are simultaneously in use in Britain, receivers will be designed so that they are capable of being switched from one system to the other. The approach to this problem is to start off with a 405-line circuit and study what circuit changes are necessary to convert it into a 625-line receiver.

The preceding survey of the two systems will have spotlighted some of the alterations needed. To summarise what is involved in this operation, the main switching operations likely to be encountered are listed below.

1. A separate u.h.f. tuner unit must be switched in.
   Alternatively a v.h.f./u.h.f. tuner may be employed.
   Both systems are in use.
2. The vision I.F. response curve must be broadened and the trap frequencies changed.
3. The sound I.F. take-off point must be moved to the 6 Mc/s intercarrier sound take-off point.
4. The sound I.F. transformers must be changed from those at 38·15 Mc/s to others at 6 Mc/s.
5. The A.M. detector must be replaced by an F.M. demodulator in the sound receiver.
6. Vision detector diode has to be reversed or replaced to maintain the same video signal polarity for the input to the video amplifier stage.
7. Possibly, an a.c. coupling network may be introduced between the detector and the video amplifier grid, instead of the normal d.c. coupling. This matter is fully discussed in Chapter 5 on Vision Detectors.
8. Video amplifier bias may have to be altered.
9. The 3·5 Mc/s inter-I.F. beat frequency rejector (if fitted) must be switched out of the video amplifier circuit. Possibly, another tuned to 6 Mc/s will be switched on.
10. The line oscillator frequency must be increased from 10,125 to 15,625.

11. Several changes are needed in the line output stage to correct width, linearity, E.H.T., boost H.T. voltage, and heater voltage for the diode rectifier which provides the E.H.T.
12. Vision a.g.c. circuit may be altered and sound a.g.c. removed.

In due course the way in which these alterations are achieved will be explained as various sections of receiver circuitry are studied.

The aim of this chapter was to arrive at a general overall understanding of the organisation of television receiver circuits. In the next two chapters a closer study of television standards and waveforms is made.

# 3

# Vision Signals

It is vitally important that a student of television engineering should become thoroughly familiar with the nature of video signals. The video signal is the electrical representation of the picture being transmitted; it contains all the information needed to re-constitute the picture at the receiver. The vision section of a television receiver is wholly concerned with processing this video signal and obeying its instructions. Quite obviously its form must be preserved carefully, since a distortion of the picture detail will alter the reproduced image whilst a distortion of the pulses will affect synchronisation.

An oscilloscope is an invaluable asset in helping an engineer to gain this essential familiarity. Both in reinforcing theoretical studies by practical experimentation, and in fault finding, nothing is more rewarding and instructive than the methodical and intelligent use of an oscilloscope. This is particularly true for all sections of the receiver after the detector, and video signal, synchronising and timebase waveforms should be studied whenever the opportunity arises. Good instruments are now available at reasonable prices; some especially designed for applications such as television. Nobody who has once learnt to appreciate its worth will ever wish to be without this powerful and pleasing tool.

As stated in Chapter 2 the video signal contains both picture detail and synchronisation pulses. These two parts of the total modulation information are required not simultaneously, but consecutively. This implies that the video signal is divided on a time basis between picture and sync. pulse information. In addition to this time division, picture and sync. information occupy different amplitude levels in the video signal. As was stated in Chapter 2, this allows for easy separation at the receiver of the two parts of the signal.

### The picture-sync. ratio

This is the ratio of the picture-detail amplitude, to the sync.-pulse amplitude. With most television systems the ratio is of the order of 7:3 for the final radiated signal, though at earlier points in the transmitter the ratio may be much different (e.g. 1:1).

The final ratio is carefully chosen and 7:3 has been found in practice to be about the optimum value. If the picture is increased at the expense of the sync. pulses, then, when the signal-to-noise ratio of the received signal falls, a point is reached when the sync. pulse amplitude is insufficient to keep the picture locked, even though the picture voltage is still of adequate amplitude to yield an acceptable picture. Conversely, if the sync. pulses are increased at the expense of picture detail, then, under similar conditions, the raster remains locked but the picture content is of too low an amplitude to set up a worthwhile result.

A choice of 7:3, or thereabouts, results in a situation such that when the signal-to-noise ratio reaches a certain low level, the sync. fails at the same time as the picture ceases to be of entertainment value. This represents the most efficient use of the television system.

**Video signal dimensions**

The time and amplitude division of a video signal is portrayed in Fig. 3.1.

Video signals are referred to in terms of 'lines'. If the left-hand leading edge of the line-sync. pulse is taken as the starting point of such a line, then the shape of the video voltage from this point to the corresponding starting point of the next succeeding line-sync. pulse, represents one *complete* 'line' of the video signal.

It will be noticed that the word 'line', when used in this context, means more than just the voltage change corresponding to the visible raster line seen on the c.r.t.; it includes the shape

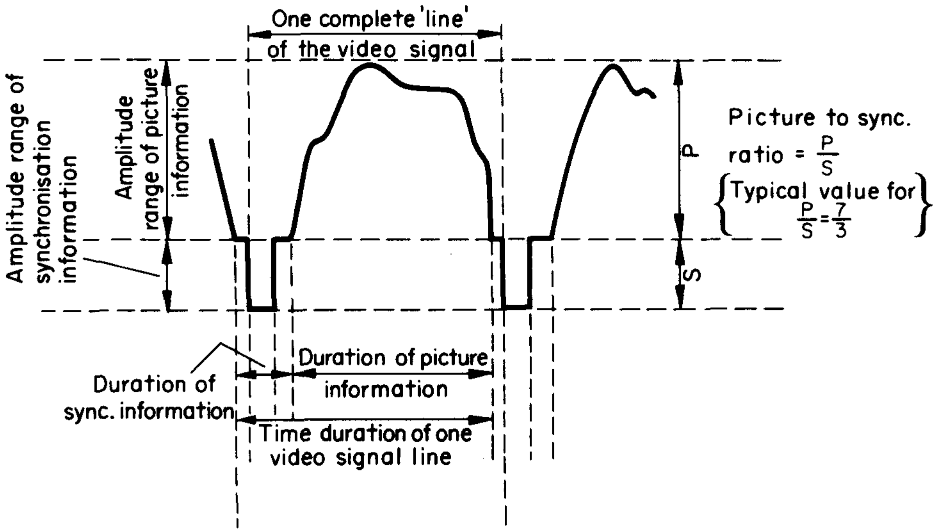



Fig. 3.1 Showing how picture and synchronisation information is divided on both a time and amplitude basis during normal picture bearing lines of a video signal

of the modulating voltage during the additional synchronisation period when the c.r.t. beam is blacked out and making its return journey. When represented graphically the horizontal or $x$-axis of a video signal has a time dimension measured in micro-seconds.

The vertical $y$-axis is scaled either in volts, or in terms of a percentage referred to the maximum amplitude level of 100%. The meaning of the terms 'positive modulation' and 'negative modulation' has been explained in Chapter 2. In the former case, as with the 405-line signal, the 100% level corresponds to 'peak-white', whilst in the latter case, as with the negatively modulated 625-line signal, the 100% level corresponds to the tips of the sync.-pulses.

The diagram of Figs 3.2 and 3.3 show details of the British 405-line and 625-line video signals. The following terms are used in describing the parts and dimensions of video signals shown.

*Line period.* This is the total time duration of one complete line of video information. With the 405-line signal there are $405 \times 25 = 10{,}125$ lines per second. The time duration of one line is therefore

$$\frac{1}{10{,}125}\text{ seconds} = \frac{10^6}{10{,}125}\ \mu\text{s} = 98{\cdot}7\ \mu\text{s} \qquad \text{(i.e. approx. 100 } \mu\text{s).}$$

Similarly, the 625-line signal consists of $625 \times 25 = 15{,}625$ lines per second, so that the line period in this case is

$$\frac{1}{15{,}625}\ \text{seconds} = \frac{10^6}{15{,}625}\ \mu\text{s} = 64\ \mu\text{s}$$

*Line blanking period.* This is the non-picture bearing part of each line during which the line synchronisation pulse is inserted. As far as the receiver is concerned it represents the time

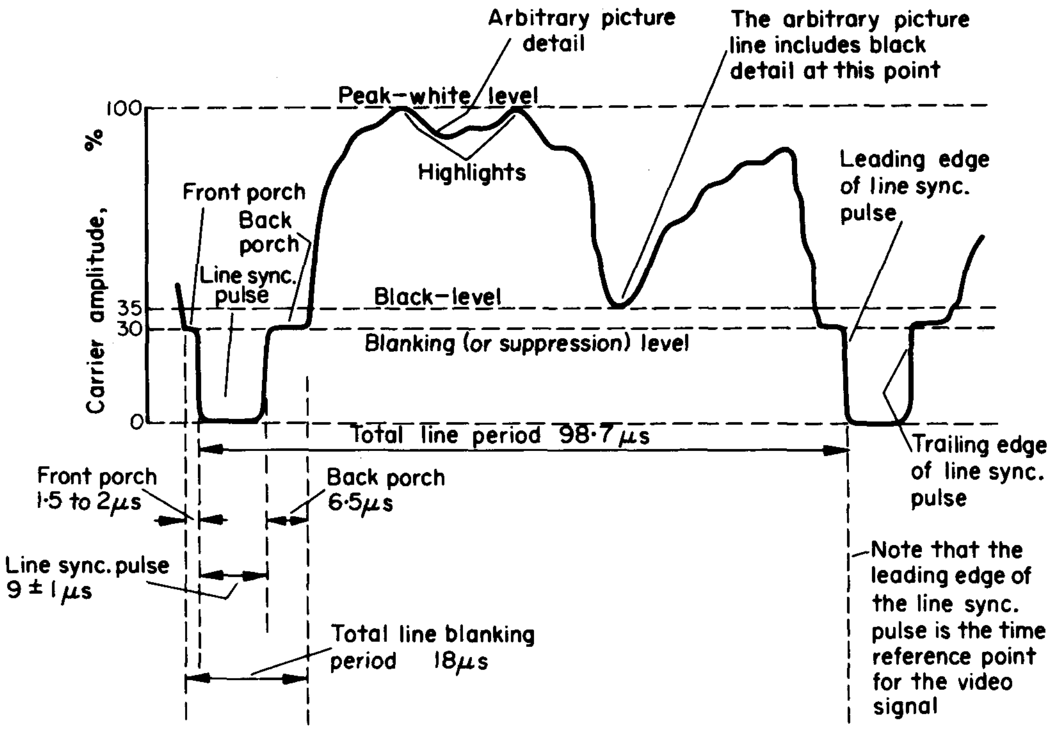



Fig. 3.2 Video signal waveform for the British positively modulated 405-line television system, showing line dimensions

allowed at the end of each picture line, for the beam flyback to be initiated and completed. This period is of 18 μs in the 405 signal and 12 μs in the 625 signal.

The line blanking period is divided into three sections. These are the front porch, the line sync. pulse and the back porch.

*Front porch.* This is a brief cushioning period at the end of each raster line, inserted between the end of the picture detail for that line and the leading edge of the line sync. pulse. Electrical circuits cannot rise or fall from one voltage level to another instantaneously. The line sync. pulses generated by the receiver in response to the incoming signal, must be regularly spaced by intervals which are precisely equal to the line duration time. For this to happen, the leading edges of the incoming line pulses must be 'seen' by the sync. separator circuit at the moment they occur.

The front porch interval allows the receiver video circuits to settle down from whatever

picture voltage level exists at the end of a picture line, to 'blanking level', before the line pulse occurs. This isolates the sync. circuits from the influence of end-of-line picture detail. The most stringent demand is made on the video circuits when peak white detail occurs at the end of a line. Despite the existence of the front porch it is possible for fault conditions at the receiver to give rise to an effect known as 'pulling on whites'. This condition is characterised by an upset in line synchronisation every time a picture line ends in white detail. The video voltage level fails to decay to blanking level before the leading edge of the line pulse occurs, and the timebase triggers late. As a result, the spot is late arriving over at the left of the screen and picture information on the next line is displaced to the left.

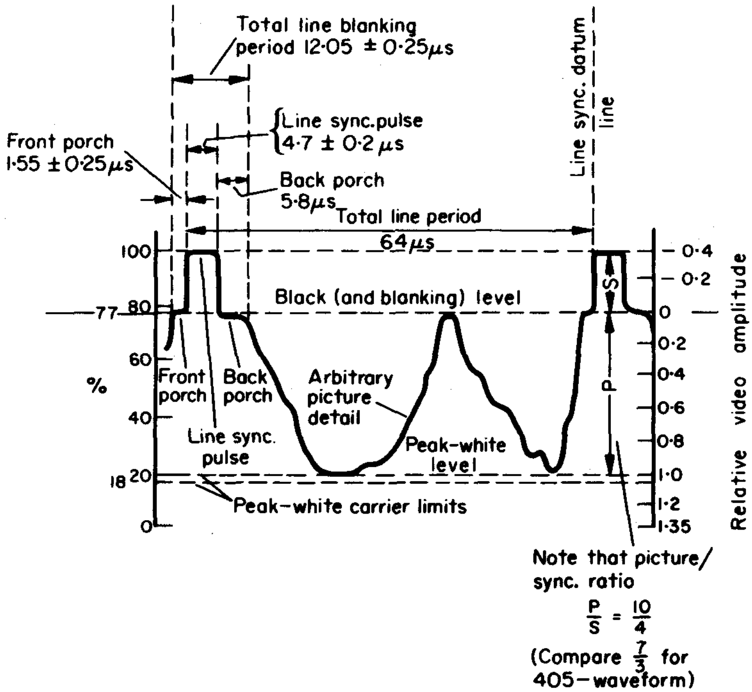



Fig. 3.3 Video signal waveform for the British negatively modulated 625-line television system, showing line dimensions

*Line sync. pulse.* This all-important pulse represents, with the corresponding field pulse sequence, the timing control of the receiver. As stated in Chapter 1, these are necessary to keep the receiver line and field timebases in precise synchronism with the distant transmitter. The nominal time durations for the line pulses of the 405 and 625-line systems are 9 μs and 4·7 μs respectively.

*Back porch.* This period allows plenty of time for line flyback to be completed before the picture detail commences again. If this were the only purpose of the back porch, however, it would not need to be as long. A most important function of this interval is to provide a regularly occurring period when the signal is held constant at a fixed amplitude equal to blanking level. It is necessary at various points in the evolution and subsequent handling of the television signal to preserve the d.c. content of the video information, and the back porch provides the necessary reference level. Whilst this is mainly a transmitter matter, the back

porch is often useful at the receiver; e.g. it can be utilised in a.g.c. circuits to produce an a.g.c. voltage which is proportional to the true signal strength and independent of picture detail.

*Peak-white level.* The level of the video signal when the picture detail being transmitted is of the maximum level of whiteness handled, is referred to as 'peak-white'. It should be noted once again that in the case of positively modulated signals the vision signal amplitude, that is to say the modulated carrier amplitude, is at its maximum (i.e. 100%) level, on peak-white. Conversely, with negative modulation, the vision signal has its minimum amplitude on peak-white. In the case of the British 625-line signal, peak-white rests at 20% of the maximum amplitude. It may be wondered why it is that peak-white is not made to take the vision signal down to zero amplitude, so using the whole available amplitude range as is done in positively modulated signals. The reason lies in the use of the 'inter-carrier' sound system used with the 625-line signal. To preserve the F.M. sound signal, both the sound and vision carriers must reach the vision demodulator circuit where they beat together to give the sound I.F. If the vision signal were made to excurse down to zero on peak-whites, the incoming vision carrier would be at zero amplitude every time peak-white occurred and this would interrupt the production of the inter-carrier sound I.F. Severe distortion would obviously result.

*Black-level.* This is the video signal level when the picture detail being transmitted is black. With the 405-line signal, black-level is at 35% of the maximum amplitude, whilst in the British 625-line signal it is at 77%.

*Blanking level.* This is the level of the signal on which the synchronisation pulses are set. With British signals, blanking level is at 30% in the 405-line case and 77% in the 625-line signal.

*The pedestal* (set-up). The difference in level (if there is one) between blanking level and black-level is called the 'Pedestal'. With the British 405-line signal a 5% pedestal is specified, although in practice it is often not used. Most other systems, including the British 625-line signal, maintain black-level as nearly as possible coincident with blanking level, i.e. there is no specified pedestal. It should be noted in this connection that blanking level is established by the synchronising and blanking circuits of the camera and is independent of the actual brightness of a scene. The camera control operator has to adjust the d.c. component of the signal so that the level corresponding to blackness in the picture is set as close to blanking level as possible (or of course to maintain the required pedestal if one is specified). The difference between black-level and blanking level is sometimes referred to as the 'set-up', and the aim in most systems is to maintain the set-up as close to zero as possible.

*Rise-time.* Electrical circuits take time to move from one voltage level to another, and what is intended to be a vertical edge to a waveform develops a slope due to this implicit delay. It is a part of the concern of the designer to minimise this effect. The rise-time of a pulse waveform is defined as the time taken for the pulse voltage to move from 10% to 90% of its maximum value. A statement of standards for a video waveform includes a specified maximum rise-time for the leading and trailing edges of the sync. pulses. This figure is normally of the order of 0·25 μs.

**Field synchronising signals**

So far the nature of the video signal during picture bearing lines has been examined. At the end of each field, a field synchronising waveform is inserted. The waveforms for the British 405-line and 625-line signals are shown in Figs 3.4 and 3.5.

The following terms are used in connection with these waveforms.

*Field blanking period.* This is the time duration of the total field synchronising waveform. During this time, as is implicit in the term 'blanking', picture information is entirely suppressed. The video signal concerns itself only with the task of initiating the flyback stroke of the receiver

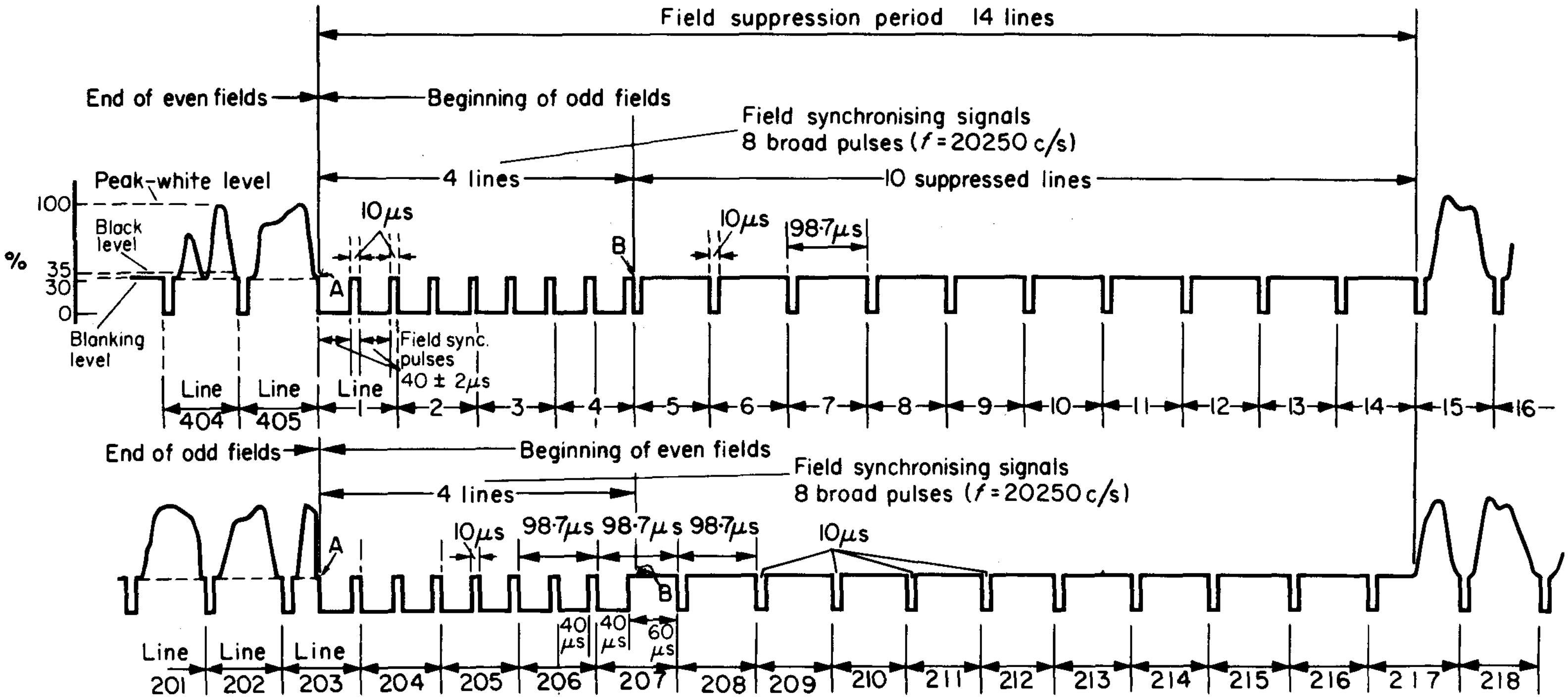



Fig. 3.4 Field synchronising waveforms of the British 405-line television signal

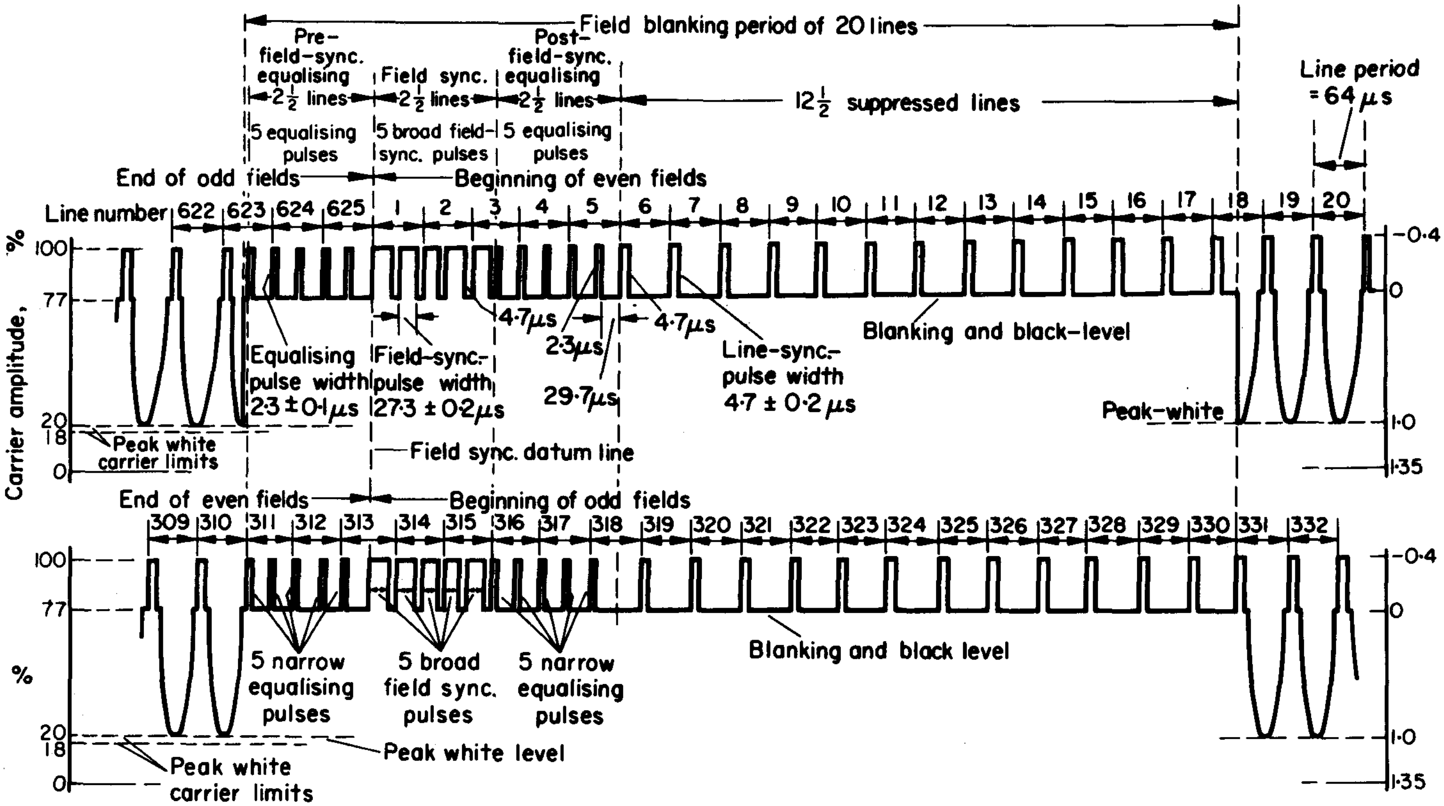



Fig. 3.5 Field synchronising waveforms of the British 625-line television signal

field timebase and of allowing time for this flyback excursion to be fully completed. Because of its length, the time dimension of the field blanking period is usually specified in terms of lines. For the 405-line signal the period is of 14 lines, whilst the corresponding figure for the British 625-line signal is 20 lines.

In both 405 and 625-line receivers the field timebase runs at the same speed of 50 c/s. It may be wondered why a period of 14 lines is allowed for field synchronisation and flyback in the one case, but 20 lines in the other. The answer is, of course, that the 'lines' are of different time duration. In the 405 case, 14 lines represents a time of $14 \times 98{\cdot}5$ µs; i.e. approximately 1400 µs. In the 625-line signal, 20 lines is a time of $20 \times 64$ µs $= 1280$ µs. The field blanking times for the two signals are thus seen to be of the same order. In the waveform inserted during this period, a number of distinct parts may be discerned.

*Field sync. pulses.* From this series of broad pulses, occurring at half-line intervals, the receiver builds a composite field sync. pulse whose purpose is to trigger the field timebase oscillator into its flyback stroke.

In the 405-line signal there are eight 40 µs pulses which occupy four lines of the 14-line field blanking period.

The British 625-line signal contains five 27·3 µs field sync. pulses which occupy $2\frac{1}{2}$ lines of the 20 line blanking period.

*Post-field-pulse suppression period.* Following the succession of field sync. pulses ample time must be allowed for the receiver field flyback movement to be completed, before picture detail is again transmitted. For this purpose the video signal is held at blanking level (i.e. the picture detail is suppressed) for an appropriate time. In the 405-line signal the suppression period is of 10 lines. With the British 625-line signal a suppression period of $12\frac{1}{2}$ lines is included, but sandwiched between this period and the field pulses is a group of equalising pulses lasting for $2\frac{1}{2}$ lines.

*Equalising pulses.* In the 625-line signal it will be noticed that there are two groups of five 2·3 µs equalising pulses inserted before and after the field pulses.

These are called the pre-field-sync. equalising pulses and the post-field-sync. equalising pulses respectively. There are no equalising pulses present in the 405-line signal. Equalising pulses are included to ensure that the composite field sync. pulse generated in the receiver has precisely the same shape on both odd and even fields. The purpose of equalising pulses will be fully understood when the chapters on differentiators and integrators, and interlacing have been studied.

*Odd and even fields.* It is necessary to distinguish between the two fields which go to make up each picture. The terms 'odd' and 'even' are used for this purpose. Unfortunately some confusion has arisen in their use, and it is important to learn the correct definition of these terms. As pointed out in Chapter 1, to set up an interlaced raster it is necessary to employ an odd number of lines per picture, so that the field timebase is triggered in the middle of the last line of one field, but at the end of the last line of the alternate field.

The field which ends in a half-line of picture information is defined as the 'odd' field, whilst that which ends in a full-line of picture information is termed the 'even' field.*

*Line numbering.* Each line of a complete picture is individually numbered. The first line of one or other of the two fields is designated line number 1, and the remainder are numbered consecutively as they occur in time and not as displayed on the c.r.t. screen.

* This follows the nomenclature used in Report No. 124 adopted by the International Radio Consultative Committee at its Plenary Assembly in Los Angeles in 1959.

In the 405-line system, the first line of the 'odd' field is labelled line number 1. This is shown in Fig. 3.4.

Having so labelled the lines it is natural that the 'odd' field should be regarded as the 'first' field of the picture, whilst the 'even' field becomes the 'second'. This appears to establish a meaning of the term 'odd', which it does not in fact have in this context.

Thus it became widely believed that the odd field was so called *because* it is the 'first' field and the alternate field named 'even' *because* it is the 'second'.

Unfortunately, this simple relationship is not true for other systems. There is nothing in the definition of the term 'odd field' which specifies that it must be regarded as the first field of a picture.

In the British 625-line system it is in fact the first line of the 'even' field which is designated as line number 1. The reason for this may be deduced from an examination of the 625 waveforms in Fig. 3.5. It is conventional in all systems to regard a field as starting from the leading edge of the first field sync. pulse in the field synchronisation waveform which triggers that field. By definition, the odd field has to be that one in which the picture detail ends half-way along a line. A study of Fig. 3.5 shows that if the line numbering were started from the beginning of the odd field, then it would in fact have to *start* in the *middle* of what is now line 313.

A little thought shows that to attempt to number the lines of the 625-line signal starting from the first odd field line would lead to an impossible situation, since in fact the first half of this *particular* line is the last half-line of the even field. This difficulty arises because the signal includes an odd number of equalising and field pulses. The matter is seen to be rather tortuous!

To sum up, however, it is only necessary to learn the definition of the term 'odd field', and then to remember, as a separate matter, that line numbering starts from the beginning of odd fields with the 405-line system, but from the start of even fields with the British 625-line signal.

*Factors determining the shape of field sync.-waveform*

It may be wondered why it is that the field sync. waveform has a somewhat complex form. The reason for this stems from the fact that the waveform has to satisfy three basic requirements. These are:

(1) It must be of a form such that at the receiver a suitable field sync. pulse may be derived from it, to trigger the field timebase.
(2) It must include provision to keep the receiver line timebase synchronised, whilst the process of the initiation and completion of the field timebase flyback stroke is taking place.
(3) It must be suitable for insertion in the middle of the last picture-bearing line of one field (i.e. the odd field), but at the end of the last picture-bearing line of the alternate (even) field.

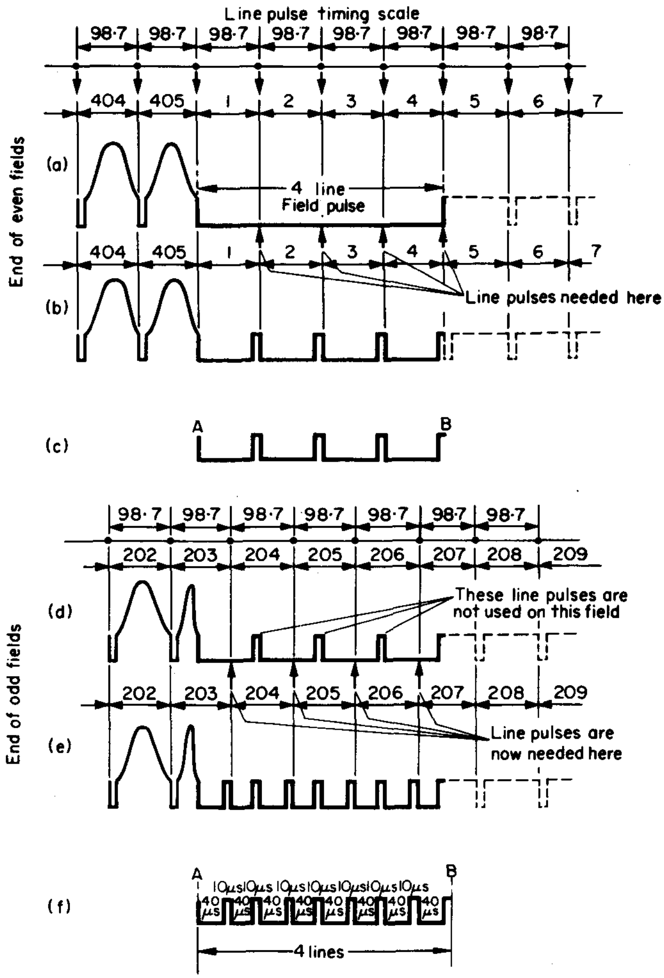
A fourth, rather more subtle requirement, which is met in the 625-line waveform but not in the 405-line case, may be stated as follows:

(4) The waveform should be such that it is technically easy to ensure that the composite field sync. pulse built at the receiver has precisely the same shape and timing on odd and even fields.

This last point is connected with interlacing. It is met by the inclusion of equalising pulses before and after the field sync. pulse sequence. The matter is discussed fully in Chapters 12 and 13.

It is instructive to build the 405-line waveform step-by-step, showing how the first three of these requirements are met.

*Step 1.* The field sync. pulse must differ substantially from line sync. pulses to allow for separation of the two types of pulses at the receiver. Both have the same amplitude but by making a field pulse much longer than a line pulse it is a simple matter to discriminate between them. In the 405-line signal the field pulse section of the field synchronising waveform is made four lines long. Fig. 3.6(a) shows a pulse of this length. Suppose this is inserted after line 405 just as it stands. It is evident that during the period of this long pulse, the line timebase will go




The basic field sync. pulse is four lines long.

Line pulses are provided by taking the signal up to blanking level 10 µs before the line pulses are needed. The trailing edge of these 10 µs pulses triggers the line timebase.

This is the waveform so far built, consisting of the original 4-line field pulse, modified to keep the line timebase in synchronism.

The modified pulse of (c) is inserted here on the alternate field. As it stands it will not keep the line timebase in synchronism.

By the inclusion of 4 more line pulses the waveform now serves this field as well as the previous one.

This final waveform is thus suitable for both fields. It is inserted, followed by 10 suppressed lines, 50 times per second (see Fig. 3.4 for the full synchronising signal).

Fig. 3.6 In these diagrams the field-sync. waveform of the British 405-line signal is built up step-by-step to show the reason for its form. The same general form will be found in almost all television signals, i.e. a succession of broad pulses interspersed with narrow ones at half-line intervals. (*N.B.* The French 819-line signal is one exception.)

off frequency. If this were allowed to happen, there would be an upset at the start of each field, since it would take some time for the line timebase to be brought back into synchronism.

*Step 2.* The line timebase must therefore be kept in synchronism, and it is necessary to interrupt the long field pulse and insert line pulses at the same intervals as they normally occur.

Thus line pulses are needed at the points marked with arrows. The technique is to take the video signal amplitude back up to blanking level 10 μs before the line pulses are needed. The waveform is then returned to zero level at the moment the line timebase needs synchronising so that the trailing edge of the 10 μs pulse serves to synchronise the line oscillator. Fig. 3.6(b) and (c) show the long pulse modified in this way.

*Step 3.* After odd fields this field pulse must be inserted in the middle of line 203. This is shown in Fig. 3.6(d). It is immediately evident that the line pulses put in to keep the line timebase running when this field pulse was used at the end of the even field, are in the wrong position to serve the same purpose in this alternate field. Line pulses are now needed at the points marked by the arrows in Fig. 3.6(d).

*Step 4.* These extra line pulses are shown inserted in Fig. 3.6(e). The waveform now built is suitable for both fields and it is seen that the original long single field pulse has now become a series of eight broad 40 μs field pulses, separated by short 10 μs pulses; the latter being spaced at half-line intervals. To allow time for the field flyback to be completed, the field pulse sequence is followed by 10 'suppressed' lines, as stated earlier. Reference back to the full waveforms of Fig. 3.4 shows the appearance of the video signal when the waveform of Fig. 3.6(f) is inserted in the middle of line 203 and at the end of line 405. The points A and B mark the start and end of the waveform built up in Fig. 3.6(f), and these same points are identified on the full diagrams of Fig. 3.4. On both fields it will be seen that the waveform is followed by a picture suppression period of 10 lines. In the case of the signal following the end of the odd field, it is obvious that as the field pulse waveform of Fig. 3.6(f) lasts an integral number of lines (i.e. 4 lines), then since this is inserted half-way through a line, it follows that it must end half-way through a line. As a result the 10-line suppression period starts half-way through line 207 and finishes half-way through line 217.

*Necessity for equalising pulses*

At this stage, it is worth noting the reason why it is desirable to include equalising pulses in a field synchronising waveform. The receiver has to separate the field sync. pulses from the line sync. pulses and process the sequence of field pulses to produce a composite field sync. pulse. For easy achievement of perfect interlacing the pulse produced must be the same on both fields.

It is easier to produce identical pulses if the waveform immediately on either side of the field sync. pulse sequence is the same on odd and even fields. This is not so in the 405-line signal. A study of Fig. 3.4 shows a considerable difference at both ends. A line sync. pulse occurs just half-a-line ahead of the first 40 μs broad field pulse after odd fields, whereas after even fields a full line separates this first field pulse from the last line pulse to occur in front of it. Looking now at the last 40 μs pulse this is separated by over half-a-line from the next succeeding line pulse on one field but on the alternate field only 10 μs separates it from the next line pulse.

The insertion of equalising pulses before and after the field pulse sequence effectively isolates the field pulses from these discrepancies. A study of the 625-line waveforms shows how these extra pulses have the effect of moving the discrepancies away from the region of the field pulses. As their name suggests, the equalising pulses have the effect of producing in the receiver field pulse processing circuitry, identical electrical conditions on odd and even fields, both before and after the arrival of the chain of field sync. pulses.

An examination of the 625-line waveforms in Fig. 3.5 shows that in other respects the waveform is similar to the 405-line waveform. There are in this case five broad field pulses in

the field pulse sequence, and these are again separated by narrow pulses at half-line intervals. In the same way, the equalising pulses are arranged at half-line intervals.

It should be noted that during the equalising and field sync. pulse sequence, it is the leading edges of the narrow equalising pulses, and the leading edges of the broad field sync. pulses, which trigger the line timebase. A comparison of the odd and even field waveforms shows that those pulses which serve this purpose on one field are not used on the alternate field. This follows logically from the discussion on the build-up of the field sync. pulse waveform.

Finally the video signal waveforms of three other television systems are illustrated to give added perspective to the subject. These are the French 819-line signal (Fig. 3.7), the Belgian 819-line signal (Fig. 3.8) and the American 525-line signal (Fig. 3.9).

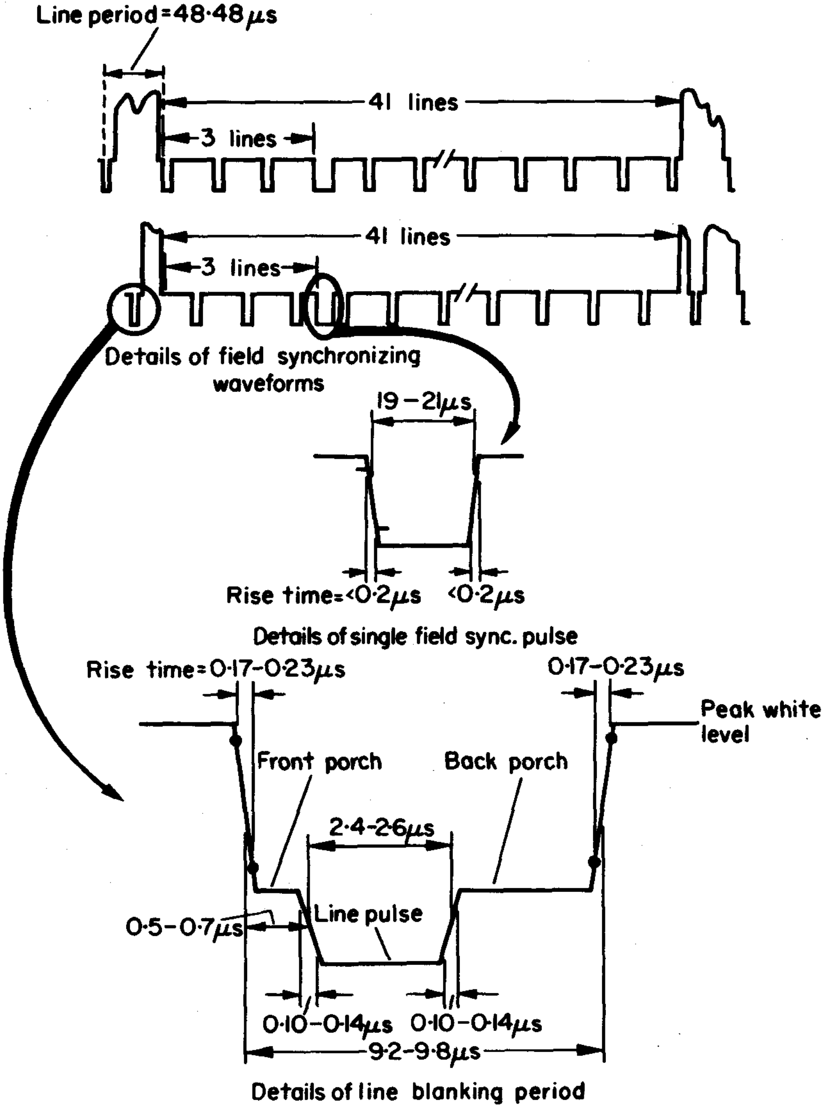



Fig. 3.7 The French 819-line video signal

The French 819-line field synchronising waveform is interesting in that it includes just one single field-sync.-pulse instead of the succession of pulses which appear in the other waveforms studied. This single pulse is, however, of 20 μs width, compared with the associated line sync. pulses which are of 2·5 μs width. Although the single field pulse is eight times as wide as a line pulse, it is none the less 'narrow' when compared with the effective field pulse width employed in the other systems.

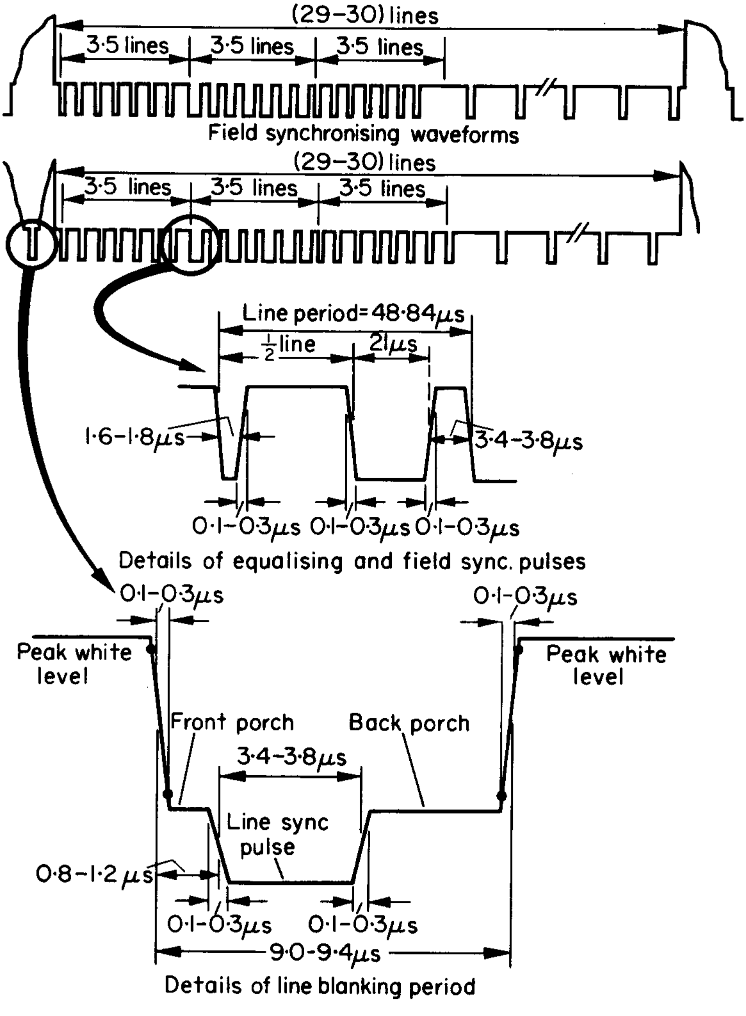



Fig. 3.8 The Belgian 819-line video signal

The total energy content of a sync. pulse generated in a receiver is roughly proportional to its width. In the French system the energy content of the derived field sync. pulse is some eight times that of a line pulse. When a succession of field pulses is used, these may be integrated (i.e. added) at the receiver to form a single pulse. Because this is so it is reasonable to regard the field pulse waveform as one long field pulse which is interrupted at half-line intervals by short duration pulses which have to be there to keep the line timebase synchronised. This was

the approach in Fig. 3.6 where the final 405-line waveform was built up from one long basic field pulse of four-lines duration. The energy content of the integrated pulse produced at the receiver may therefore be as much as thirty times that of the 10 μs line pulses.

The ratio of field pulse to line pulse energy is therefore much less for the French system

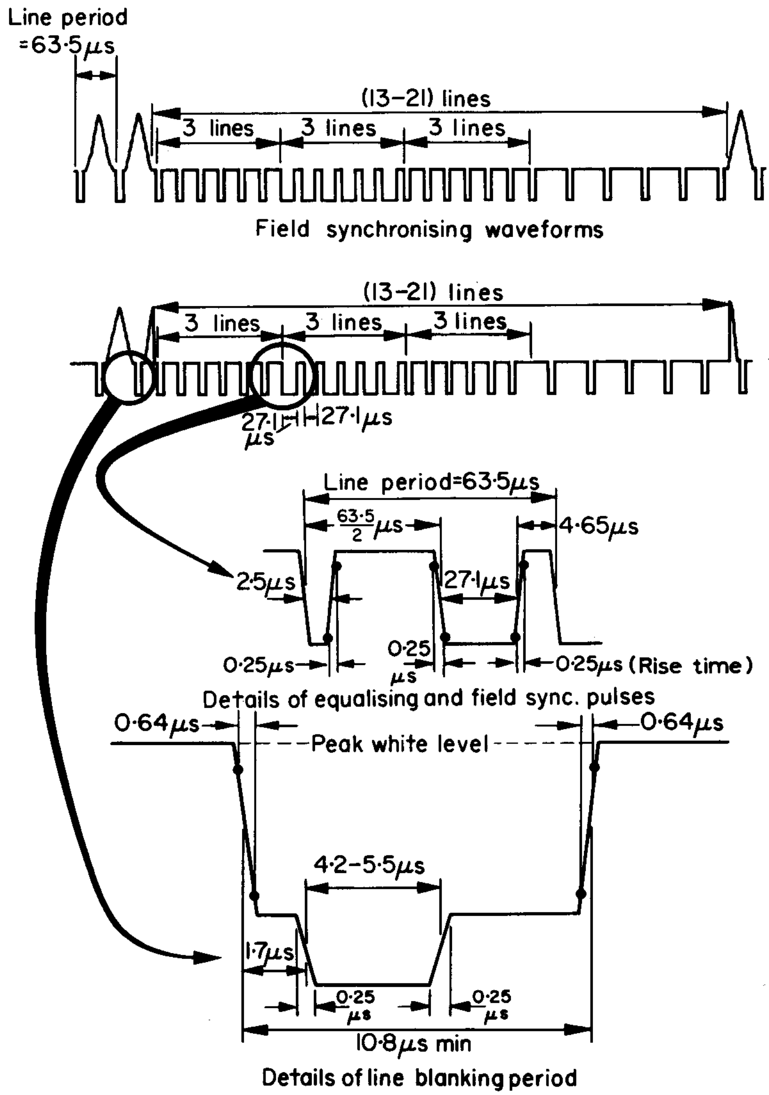



Fig. 3.9 The American 525-line video signal

than for the others studied. Too much must not be read into this comparison, however, since it is not true to say that the longer the field pulse the better the result. The essential requirement is that the receiver circuitry shall be able to discriminate between field and line pulses and the two types of pulses must merely be sufficiently different to allow for efficient sorting.

The Belgian 819-line signal is seen to differ from the French 819-line signal. There are seven broad field-sync. pulses in the Belgian signal, and these are preceded and followed by groups of seven equalising pulses. This compares with the single field sync. pulse and no equalising pulses of the French signal. The line blanking periods are similar in the two cases.

In the American 525-line video signal six broad field-sync. pulses are employed, and six short equalising pulses appear before and after the field-sync. pulses.

Referring back to the British 625-line video signal shown in Fig. 3.5, it will have been noticed that this is drawn with the sync. pulses 'pointing upwards'. This was done so that the appropriate modulation amplitudes for the various main levels of the signal could be shown alongside the diagram. With negatively modulated carriers, as has been described previously, the sync. pulse tips correspond to maximum carrier amplitude. It is therefore instructive to draw the video signals to show the envelope shape of the modulated carrier. However, it allows for an easier comparison of the signals employed by different systems if they are all drawn the same way up, and the last three waveforms shown are presented in this way. Of these three systems, the French and Belgian use positive modulation whilst the American 525-line signal is negatively modulated.

The C.C.I.R. 625-line and the Australian 625-line video signals are basically the same as the British 625-line signal already examined. The British system employs rather different sideband characteristics and a slightly greater bandwidth. These matters are discussed in the next chapter.

# 4

# Television Signal Standards

Standards are specifications of the technical details of a television system. A complete statement of standards covers every aspect of a system, with a full description of the transmitted signals on sound and vision. This includes such details as the types of modulation used on sound and vision; sideband characteristics and bandwidths occupied; time and amplitude dimensions and tolerances of video waveforms, etc., etc.

Some of these standards were studied in the previous chapter, which dealt with the nature of the video signal waveforms that form the modulation information carried by the vision signal. This chapter deals with the specifications of complete radiated television signals.

As a starting point, the information shown in Table 4.1 should be studied carefully. Here details of five of the principal monochrome television systems are listed side by side for purposes of comparison. Much of this information needs no further explanation since its significance has been explained in earlier chapters. The remaining matters are now discussed.

### Aspect ratio

This is the ratio of the width to the height of a television picture. An aspect ratio of 4:3 has been universally adopted for television systems. The picture is made wider than it is high, because the majority of movement is in the horizontal plane. The actual ratio of 4:3 was chosen in the first place to bring the television picture into line with the cinema picture film frames which already had these proportions.*

### Positive and negative modulation

The meaning of these terms has been explained. It is perhaps worth summarising the main factors which govern the choice between the two methods. These are:

(1) The effect upon the picture of impulsive interference.
(2) The effect upon synchronisation of impulsive interference.
(3) The peak power available from the transmitter.
(4) The ease with which efficient a.g.c. is available at the receiver.

(1) The effect of interference will be fully treated later in the work. Briefly it may be said that as far as the picture is concerned, the advantage is with negative modulation, where impulsive interference produces *black* spots. With positive-modulation larger and more trying white blobs are produced. Limiting circuitry is required to lessen the effect.

(2) Against this, impulsive interference is much more troublesome in its impact upon synchronisation in negative than in positive modulation receivers. This is because impulsive

* It should be noted that the rectangular display area of the c.r.t. in television receivers often has an aspect ratio which is more nearly 5:4 than 4:3.

interference causes large steep-fronted random *increases* in signal amplitude, and since synchronisation information is present at maximum carrier amplitudes it is more difficult to eradicate the effects of the spurious interference pulses.

(3) With positive modulation, picture highlights correspond to maximum carrier amplitude. Should the drive to the modulated R.F. amplifier be too high, the picture suffers distortion when peak-white detail is being handled. For this reason it is necessary to limit the peak power developed by a given transmitter to a pre-determined maximum level.

**Table 4.1**

*Vision and sound signal standards for the principal monochrome television systems*

| Signal standard | British 405 | British 625 | European C.C.I.R. 625 | French 819 | American 525 |
|---|---|---|---|---|---|
| Interlace ratio (i.e. No. of fields per picture) | 2:1 | 2:1 | 2:1 | 2:1 | 2:1 |
| No. of lines per picture | 405 | 625 | 625 | 819 | 525 |
| Line frequency (lines per second) | 10,125 | 15,625 | 15,625 | 20,475 | 15,750 |
| Line period (μs) | 98·7 | 64 | 64 | 48·84 | 63·5 |
| Field frequency (fields per sec) | 50 | 50 | 50 | 50 | 60 |
| Picture frequency (pictures per sec) | 25 | 25 | 25 | 25 | 30 |
| Aspect ratio | 4:3 | 4:3 | 4:3 | 4:3 | 4:3 |
| Sense of modulation | positive | negative | negative | positive | negative |
| Black level as % of peak carrier | 35% | 77% | 75% | 25% | 75% |
| Blanking level as % of peak carrier | 30% | 77% | 75% | 25% | 75% |
| Peak white level as % of peak carrier | 100% | 20% | 10% minimum | 100% | 15% |
| Attenuated (i.e. vestigial) sideband | upper | lower | lower | upper | lower |
| Highest video modulation frequency (c/s) | 3 Mc/s | 5·5 Mc/s | 5 Mc/s | 10·4 Mc/s | 4 Mc/s |
| Sound carrier modulation | A.M. | F.M. | F.M. | A.M. | F.M. |
| Sound carrier deviation | — | ±50 kc/s | ±50 kc/s | — | ±25 kc/s |
| Sound pre-emphasis (μs) | — | 50 μs | 50 μs | — | 75 μs |
| Position of sound carrier with reference to vision carrier | −3·5 Mc/s | +6 Mc/s | +5·5 Mc/s | −11·15 Mc/s | +4·5 Mc/s |
| Channel bandwidth | 5 Mc/s | 8 Mc/s | 7 Mc/s | 14 Mc/s | 6 Mc/s |

In the case of negative modulation, it is the synchronisation pulses which exist at maximum carrier amplitudes. Hence any non-linearity resulting from driving the transmitter harder, has no serious consequences since synchronisation pulses are not so adversely affected by non-linearity. It is estimated that a given transmitter may be allowed to develop up to 30% more peak power when employed to handle negative rather than positive modulation.

(4) As far as a.g.c. is concerned, since synchronisation pulses are at maximum carrier amplitude, it is possible to monitor this amplitude and derive an a.g.c. voltage which is truly dependent upon the strength of the incoming signal. In the case of positive modulation, the instantaneous maximum carrier amplitude depends not only upon the strength of the signal, but also upon the nature of the picture modulation. However, effective a.g.c. circuitry is available for both systems, so that this cannot be regarded as a vital issue.

A study of the television systems currently in use shows that negative modulation is more favoured than positive.

**Sound modulation**

The systems using F.M. employ the inter-carrier sound technique described in Chapter 2. The system has the advantage that the sound I.F. must always be correct, and the quality of the sound is virtually independent of local oscillator tuning. The amplification of sound and vision signals through a common I.F. amplifier chain represents a further advantage. The sound demodulator circuitry is more complex however. Also, since the vision is an A.M. signal and the sound is F.M., cross modulation between the two signals is less of a problem.

*Sound carrier deviation.* This figure, applicable only to F.M. signals, indicates the maximum extent of the sound carrier frequency swing, corresponding to modulation by the loudest signal accommodated.

*Pre-emphasis.* This device is used to enhance the signal-to-noise ratio of an F.M. signal. A study of the way in which noise power is distributed in an F.M. signal shows that it results in more interference at higher audio frequencies than lower.

At the transmitter the audio frequency signal is given a rising characteristic before application to the modulator. This favourable treatment of higher frequencies is called pre-emphasis. At the receiver, in order to restore the audio frequency components to their correct relative amplitudes, it is necessary to pass the audio signal through a de-emphasis filter having a downward slope equal to the upward slope employed at the transmitter. In attenuating the higher frequencies in this way, the noise present in the signal is also attenuated. The result is that the higher audio frequency components finish up with their normal correct amplitudes, but the noise power is decreased to a level which is less than it would otherwise be.

The degree of pre-emphasis is stated in terms of the time-constant of a simple filter circuit whose slope equals that of the transmitted rising audio frequency characteristic.

To de-emphasise it is only necessary to shunt the audio signal path at the receiver with a capacitor-resistor network having this same time-constant.

**Vestigial sideband signals**

As is well known, when a radio frequency carrier wave is amplitude modulated, side-frequencies are produced on each side of the carrier frequency. Each single frequency component of the modulating signal produces its own pair of side-frequencies. Those side-frequencies situated above the carrier are collectively called the Upper Sideband and those below form the Lower Sideband.

Since a single modulating frequency $f_m$ produces the two side frequencies $f_c+f_m$ and $f_c-f_m$,

where $f_c$ is the carrier frequency, it is obvious that the frequency space occupied by an A.M. transmission must be twice the frequency of the highest frequency component of the modulating signal.

The frequency components present in the video signal which modulates the vision carrier, extend from 0 c/s (i.e. 'd.c.') up to several megacycles. For example, as shown in Table 4.1 the highest modulating frequencies used in the British 405 and 625-line signals are 3 Mc/s and 5·5 Mc/s respectively, whilst the corresponding figure for the French 819-line system is 10·4 Mc/s. It follows that double sideband *vision* signals in these cases would occupy 6 Mc/s, 11 Mc/s and 20·8 Mc/s respectively.

The actual bandspace allocated to the television channels would have to be greater still. It is not possible to terminate the bandwidth of a signal abruptly at the edges of the sidebands. The attenuation slope usually occupies a matter of approximately 0·5 Mc/s. This adds 1 Mc/s to the required total bandspace. In addition to this each television channel has its associated sound signal, the carrier frequency of which is situated just outside either the upper or lower limits of the vision signal. This adds a further 0·25 Mc/s to the channel width necessary, so that practical figures for the transmissions discussed would be 7·25 Mc/s, 12·25 Mc/s and 22·05 Mc/s.

The modulating information borne by an A.M. wave is, however, fully contained in *each* sideband and provided the carrier frequency is also present, one sideband may be suppressed altogether. Thus the modulating frequency $f_m$ is obtainable from the upper sideband by taking $(f_c+f_m)-f_c$ or from the lower sideband by taking $f_c-(f_c-f_m)$. This principle is fully exploited in single sideband techniques which represent an absorbing study in their own right.

It is not possible, however, to go to this extreme of fully suppressing one complete sideband in the case of a television signal. This is because the video modulation includes frequency components which extend from the high upper limits of several Mc/s right down through very low frequencies to d.c. information. It is impracticable to design a filter which will cut out one complete sideband and leave unscathed the carrier and the low frequency components of the other sideband. As a compromise, a part of one sideband is suppressed. The radiated signal then consists of one complete sideband, together with the carrier and the 'vestige' (or remaining part) of the partially suppressed sideband.

Fig. 4.1 illustrates the saving of bandspace which results from this technique. The diagram contrasts the appearance of the vestigial sideband British 625-line vision channel, with a double sideband signal using the same standards.

The vestigial sideband *vision* signal is seen to occupy a bandwidth of 7·75 Mc/s. The complete television channel with its associated sound transmission, is allocated a bandspace of 8 Mc/s. This represents a bandspace saving of 4·25 Mc/s per channel, when compared with the 12·25 Mc/s space which would be required by the corresponding double sideband signal.

A study of the vestigial sideband signal characteristic shows that the upper sideband is fully radiated but in the lower sideband only frequencies up to 1·25 Mc/s are included. As with the upper edge, the end slope has a width of 0·5 Mc/s, so that attenuation of the lower sideband starts at 1·25 Mc/s and full suppression is achieved at 1·75 Mc/s.

Table 4.2 summarises the frequency dimensions of the vestigial sideband vision signals of various television systems, and in Fig. 4.2 these television channels are represented diagrammatically. The frequency axis is in each case scaled relative to the vision carrier, which is marked as 0 Mc/s. This makes the diagrams very informative since detail such as the widths of the upper and lower sidebands and the relative positions of the sound carriers is easily read off. It will be observed that the sound carrier is always positioned as far away from the vision

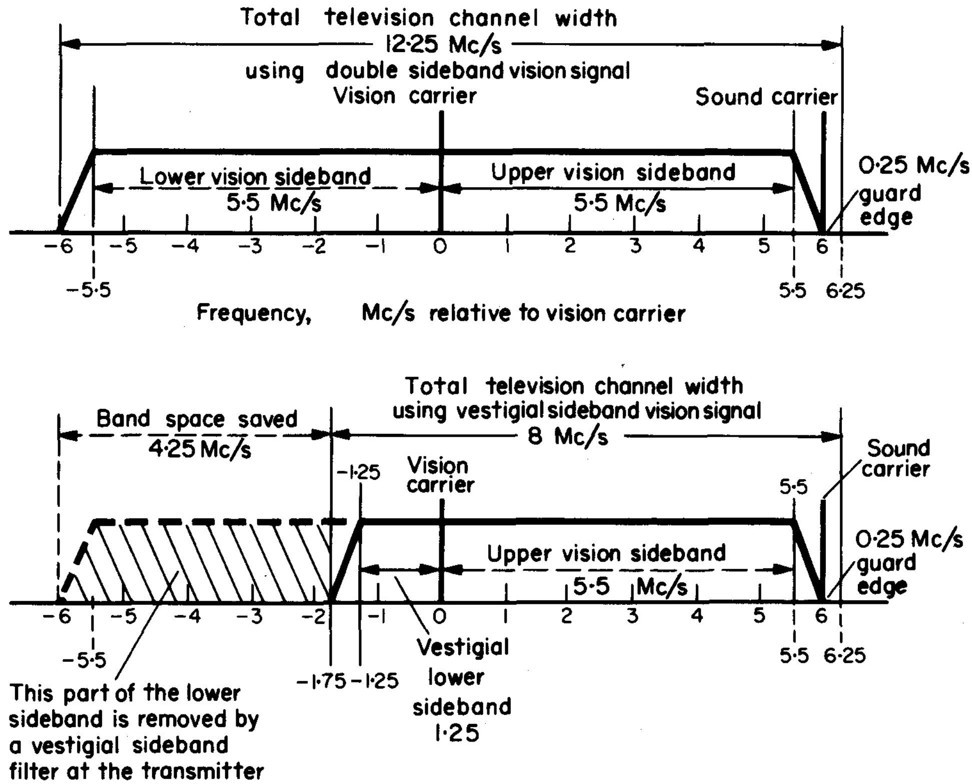



Fig. 4.1 Diagrams illustrating the saving of bandspace which results from the use of 'vestigial' sideband rather than double-sideband vision signals. The standards shown are those of the British 625-line signal

**Table 4.2**

*Vision signal sideband standards for some of the principal monochrome television systems*

| Radiated signal sideband standard | British 405 | British 625 | European C.C.I.R. 625 | French 819 | American 525 |
|---|---|---|---|---|---|
| Fully radiated sideband | lower | upper | upper | lower | upper |
| Width of flat section of full sideband (=maximum modulating frequency) | 3 Mc/s | 5·5 Mc/s | 5 Mc/s | 10·4 Mc/s | 4 Mc/s |
| Width of end-slope of full sideband | 0·5 Mc/s | 0·5 Mc/s | 0·5 Mc/s | 0·75 Mc/s | 0·5 Mc/s |
| Attenuated (i.e. vestigial) sideband | upper | lower | lower | upper | lower |
| Width of flat section of attenuated sideband | 0·75 Mc/s | 1·25 Mc/s | 0·75 Mc/s | 2·0 Mc/s | 0·75 Mc/s |
| Width of end-slope of attenuated sideband | 0·5 Mc/s | 0·5 Mc/s | 0·5 Mc/s | 0·75 Mc/s | 0·5 M/cs |

carrier as possible; i.e. at the extremity of the fully radiated sideband. This is the logical place for it since it makes for minimum interference between the two signals.

Both vision and sound carriers beat with the same local oscillator frequency at the receiver to produce I.F.s which are separated by the same frequency difference as are the carrier frequencies. By placing the sound carrier at the edge of the fully radiated sideband, the inter-I.F. beat frequency component produced at the vision detector is just outside the passband of the video amplifier and can therefore be rejected from the video signal path.

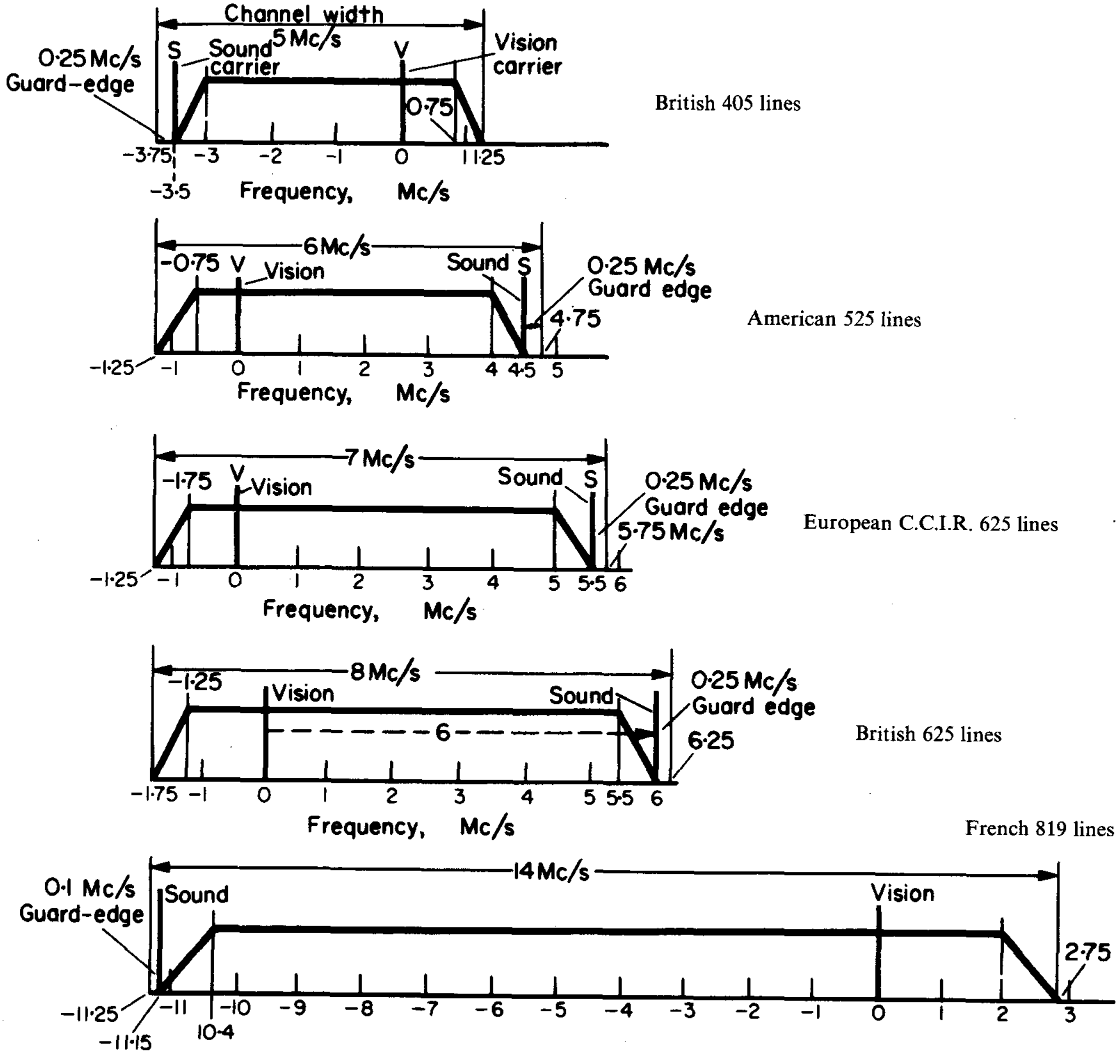



Fig. 4.2 Television channels showing standards employed by the British (405 and 625), American, European (C.C.I.R.) and French signals

*Note* that a 'guard edge' is allowed in each case, to the side of the sound signal.

The lower frequency edges of all the signals have been drawn vertically between one another to draw attention to the differences between the bandwidths occupied.

The diagrams show that a *guard-edge* is allowed on the sound carrier side of the television channel to allow for adequate inter-channel separation. The actual sound signal bandwidth is of course too small to show on these diagrams.

The diagram in Fig. 4.3 shows how two British 625-line adjacent channels are disposed in Band IV.

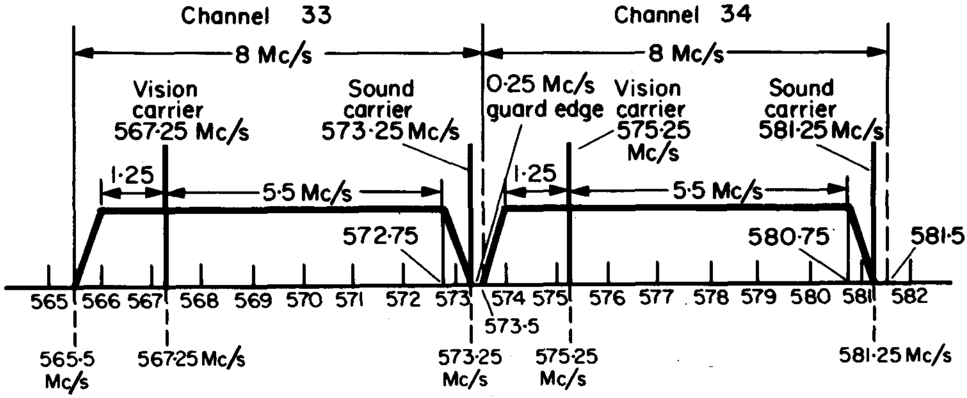



Fig. 4.3 Showing how the adjacent British 625-line vestigial sideband television channels fit into Band IV

*Reception of vestigial sideband signals*

The fact that some modulation frequencies are present in both sidebands of the vestigial signal, whilst those above a certain limit are present in one sideband only, must be taken into account at the receiver. If the shape of the vision receiver frequency response curve were identical in form to that of the vision transmitter, then all sideband frequency components would be equally treated. When fed to a perfect detector those components present in both sidebands would give rise to a total of twice the output voltage from the detector as that produced by components present in one sideband only. This means that if the vision carrier were successively modulated to an equal depth by a series of frequencies throughout the video frequency range employed by the system, and the resulting voltage output from the detector recorded, the output voltage against input frequency characteristic obtained would have the form shown in Fig. 4.4.

Two examples are shown. In the case of the 405-line signal, the vestigial sideband extends to 0·75 Mc/s above the carrier. Thereafter this sideband is linearly attenuated down to zero at 1·25 Mcs. The detector voltage output would thus be twice as great between 0 c/s and 0·75 Mc/s than between 1·25 Mc/s and 3 Mc/s. Between 0·75 Mc/s and 1·25 Mc/s the output voltage would fall off linearly following the sideband attenuation slope of the transmitter.

Similarly, in the 625-line case, in which the vestigial sideband extends to 1·25 Mc/s, the detector output voltage would be twice as great between 0 Mc/s and 1·25 Mc/s than between 1·75 Mc/s and 5·5 Mc/s. Frequencies between 1·25 Mc/s and 1·75 Mc/s, would give rise to intermediate levels of output corresponding as before to the sideband attenuation slope.

To correct this discrepancy it is necessary to so shape the receiver response curve that the frequencies present 'twice' are afforded less amplification than those occurring in one sideband only. This is easily arranged, as is demonstrated in the diagrams of Fig. 4.5.

The response curve is shaped to place the vision carrier half-way down the side corresponding

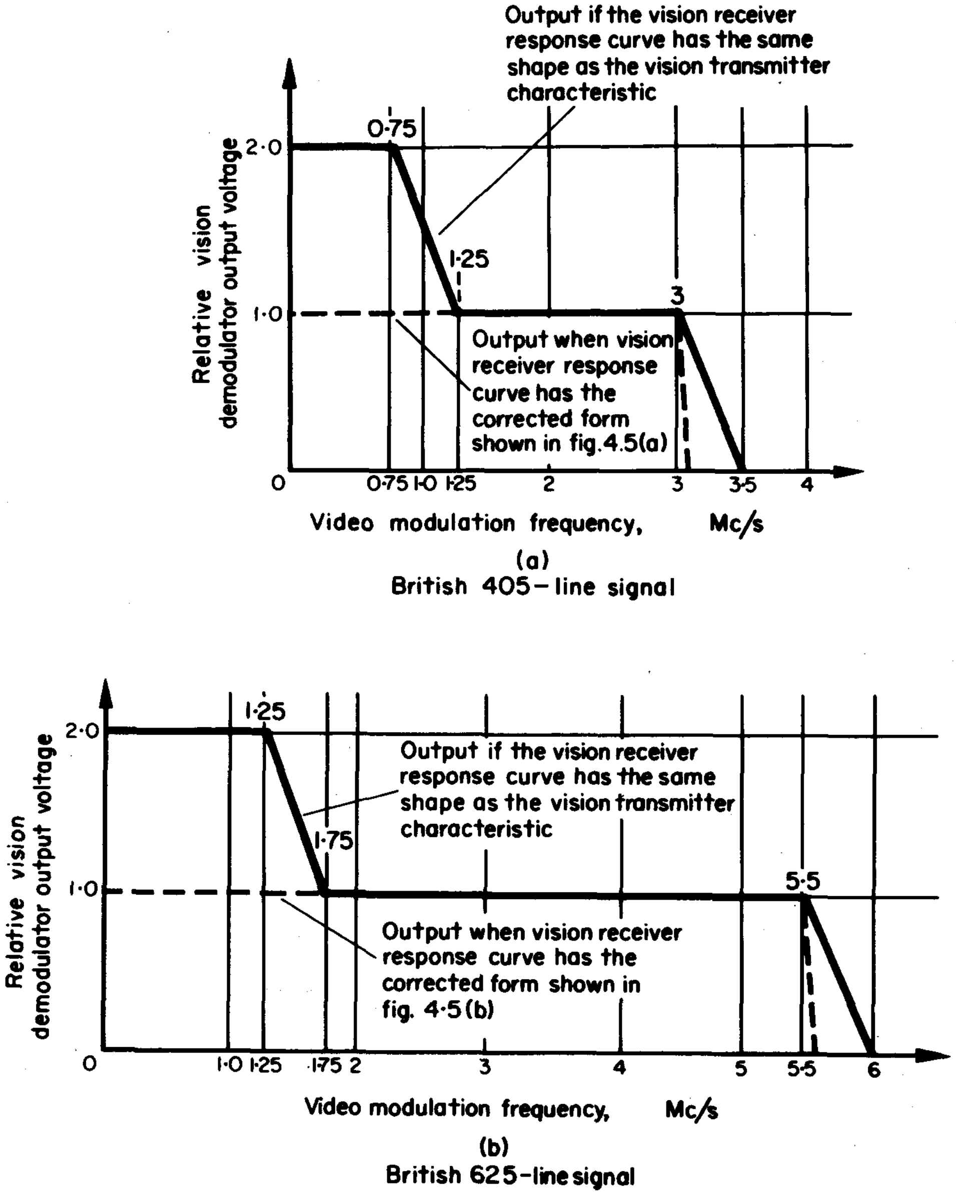
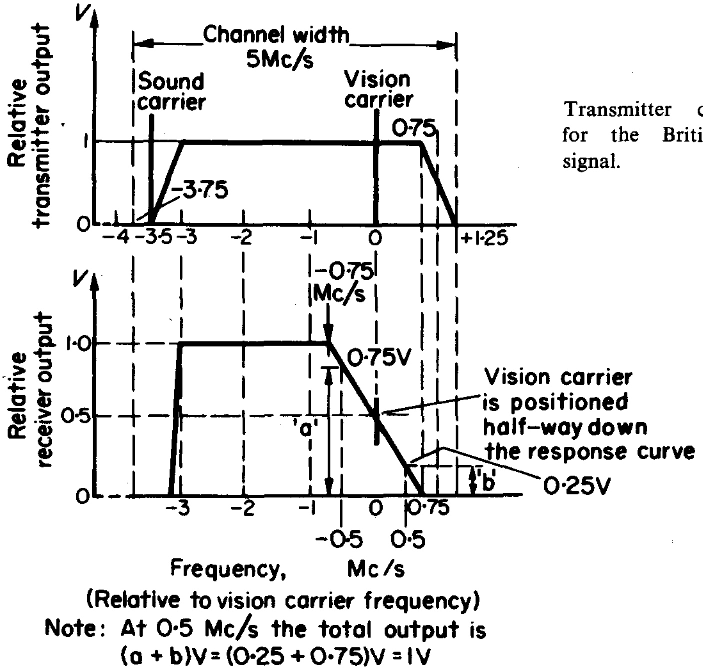



Fig. 4.4 Receiver vision demodulator output-voltage/input-frequency characteristics illustrating the need for the specially shaped vision receiver response curves used with vestigial sideband signals

to the suppressed sideband. The width of the sloping edge on which the carrier is positioned is twice the width of the vestigial sideband.

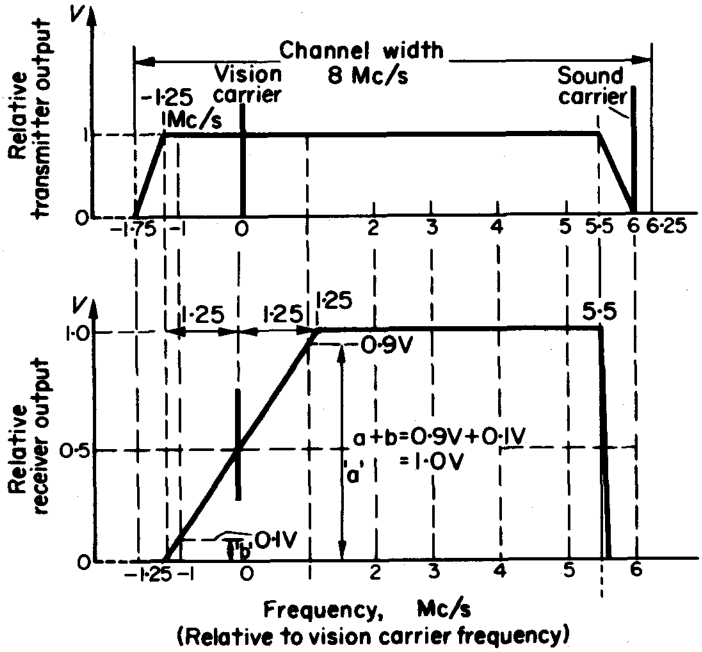
To understand how this achieves the required result, it is best to study the diagrams and consider the treatment afforded to various frequencies within the video bandwidth. In the diagram for the 405-line signal, frequencies between 3 Mc/s and 0·75 Mc/s, i.e. those present




Transmitter characteristic for the British 405-line signal.

Required vision receiver characteristic if all modulation frequency components are to give equal outputs.

Fig. 4.5(a) Transmitter and vision receiver characteristics for the British 405-line signal




Transmitter characteristic for the British 625-line signal.

Vision receiver characteristic for the British 625-line signal.

Fig. 4.5(b) Transmitter and vision receiver characteristics for the British 625-line signal

in the lower sideband only, are seen to give unit output. Suppose a component at 0·5 Mc/s is now considered. This is present in both sidebands. A study of the diagram shows, however, that the total detector output at this frequency is again unity. The component in the lower sideband gives rise to an output of $a$ volts, whilst that in the upper sideband gives rise to $b$ volts. From the geometry of the figure it is obvious that $(a+b)=1$.

As a further example consider the response to a frequency of 1 Mc/s in the 625-line case. This component in the vestigial sideband gives rise to an output of 0·1 V, whilst in the upper sideband, it gives rise to an output of 0·9 V. Again the sum of the two is unity, so that the same output is achieved as for frequencies between 1·25 Mc/s and 5·5 Mc/s, which occur in the upper sideband only.

It is interesting to note from Table 4.2 that the majority of television systems have a vestigial sideband which is only 0·75 Mc/s wide. The British 625-line signal is 1·25 Mc/s wide, whilst the French 819-line signal is 2 Mc/s wide. One advantage of a rather wider vestigial sideband (sometimes called an extended vestigial sideband) is that the tuner-unit oscillator control becomes less critical. This is demonstrated by comparing the British 405-line and 625-line signals. In the 405-line case shown in Fig. 4.5a the oscillator has to be tuned to position the carrier mid-way down a response curve slope which is 1·5 Mc/s wide. The corresponding slope for the 625-line receiver shown in Fig. 4.5b is 2·5 Mc/s wide. For a given amount of oscillator mistuning (or perhaps 'drift', after initial correct tuning), the degradation of the picture quality is less in the latter case.

**Highest video modulation frequency**

It is necessary to understand some of the factors which lie behind the determination of the highest video frequency used with a given system. Reference to Table 4.1 shows that the greater the number of lines employed by a system, the higher this figure becomes. The matter is important because upon it depends the maximum fineness of picture detail which can be resolved.

*Vertical resolution.* In the vertical plane the picture is divided into a finite number of lines. It must be remembered that not all the radiated 'lines' bear picture information (i.e. are active lines). The British 625-line signal has a field suppression period of 20 lines, so that a total of 40 lines per picture are suppressed. This leaves $(625-40)=585$ active lines per picture. The corresponding figure for the 405-line signal, where the field suppression period is of 14 lines, is $(405-28)=377$ active lines per picture.

The ability to resolve picture detail in the vertical plane is referred to as *vertical resolution.* Quite clearly, the greater the number of lines used, the better the vertical resolution. To illustrate this, and to study the whole question of quality of resolution, the best approach is to imagine that the picture to be transmitted consists of a chequer-board pattern of black and white squares.

Since the aspect ratio is 4:3, suppose, as a starting point, that a simple pattern consisting of $4\times 3=12$ squares is to be resolved using the British 625-line system. There are then four vertical columns each of three squares, so that with the 625-line signal, 585 lines are shared between the three squares in a given column; i.e. 195 lines pass through each square. Alternate groups of 195 lines in a given column will be white and black.

If the number of squares in each direction is now doubled, there are 8 columns of 6 squares, and approximately 98 lines pass through each square. As the number of squares is increased, the number of lines available to resolve each square diminishes. Obviously, the maximum number of squares in a vertical column which the system can resolve equals the number of

active lines present. In this limiting case one line resolves one square in each column, and the maximum number of squares possible in a vertical column is 585; alternate squares being coloured black and white.

If the number of squares were increased further than this, the raster lines would be thicker than the individual squares and the pattern could not be resolved.

*Horizontal resolution.* Nothing has so far been said about the problem of resolving the squares in the horizontal plane.

The argument so far has established that the maximum number of squares (picture elements) which can be resolved along a vertical line drawn through the picture, is 585. A chequerboard pattern having 585 squares per vertical column, and an aspect ratio of 4:3 would have $(585/3) \times 4 = 780$ squares along a horizontal line.

It will be observed that this pattern could be arrived at by drawing vertical lines across the raster, of equal density to the horizontal raster lines. For the purpose of the discussion adjacent lines are assumed just to touch. An all-over pattern of tiny squares (often referred to as elements) results. The maximum demand upon the system occurs when alternate squares are black and white. To resolve this pattern requires equal horizontal and vertical resolution, which is an obvious aim to seek, in designing a television system. The pattern is illustrated in Fig. 4.6.

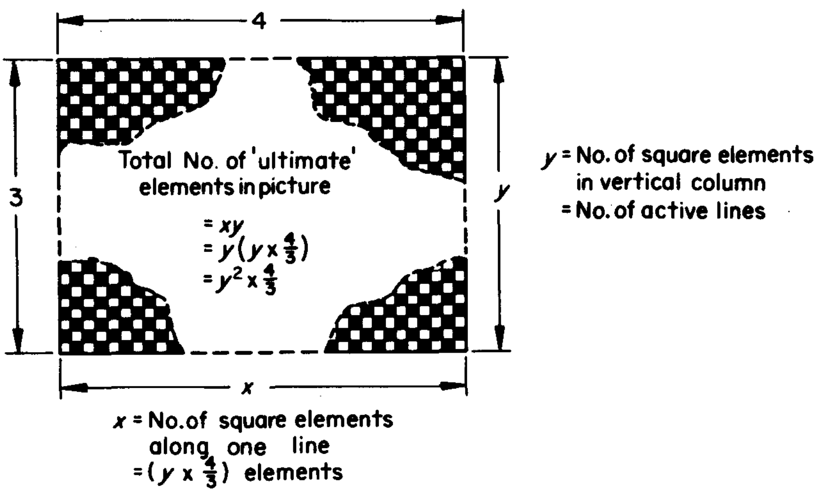



Fig. 4.6 Illustrating the chequerboard method of studying vertical and horizontal resolution For the British 625-line signal $y = 585$ elements and $x = 780$ elements.

The vertical definition, as has been seen, is dictated by the number of lines used in a system. It remains to examine what factors determine the ability of the system to give an equal degree of horizontal resolution.

In the case discussed, along each horizontal line there are 780 squares which are alternately black and white. To resolve them the scanning spot on the receiver c.r.t. screen, must change alternately from black level to peak white level. The intensity of the scanning spot is controlled by the video signal which modulates the beam current. It follows that the video signal must switch continuously along each line, between the voltage levels corresponding to black and peak white. Each adjacent pair of one black square and one white square constitutes one complete cyclic change. Along one line there are $780/2 = 390$ complete cyclic changes.

To set up the pattern postulated therefore, the picture signal waveform must be a periodic square wave and between two successive line blanking periods on the video signal there must be 390 complete square wave cycles. It remains to translate this into terms of frequency; i.e. to calculate how many times per second the video modulation amplitude must change in order to resolve this detail.

This is easily arrived at because the length in μs of one complete visible line on the c.r.t. screen is known. The total duration of one line in the 625-line system is 64 μs. The line blanking period lasts 12 μs, which means that the active, or picture bearing, part of each line lasts for $(64-12)=52$ μs. Hence along a visible raster line of length 52 μs, there are 390 complete square waves. The time duration of one square wave (i.e. its periodic time) is thus $52/390$ μs $=0{\cdot}1333$ μs. The frequency of a wave is simply the reciprocal of its periodic time. Since there are $10^6$ μs in 1 second, and each cyclic change takes 0·1333 μs, it follows that these changes are taking place at the rate of $10^6/0{\cdot}1333$ c/s, i.e. at a frequency of approximately 7·48 Mc/s.

During each active picture line, therefore, the vision carrier amplitude must vary between black level and peak white, at a frequency of 7·48 Mc/s.

It must not be forgotten that it is square waves which have been discussed. All complex periodic waves may be shown to consist of the *sum* of a series of individual sine wave components, one (the fundamental) at the frequency of the periodic wave, and the others at a succession of harmonic frequencies of increasing order of magnitude. A square wave, which is an example of a complex wave, may be analysed and shown to be equivalent to the algebraic sum of the amplitudes of a whole series of sine waves consisting of the fundamental, plus an infinite series of *odd* harmonics. The importance (i.e. the amplitudes) of these harmonics, diminishes as the order of harmonic increases, and a first approximation to a square wave may be arrived at by adding components up to the 11th harmonic, and neglecting the remainder.

This means that to reproduce a square wave of frequency 7·48 Mc/s the electrical circuits would need to be capable of handling signals at least up to $11\times7{\cdot}48=82{\cdot}28$ Mc/s. This is grossly excessive, and would necessitate an impossible signal bandwidth.

As an attainable compromise, a sine wave response to this fine square wave chequerboard pattern is accepted. This is illustrated in Fig. 4.7. Instead of the scanning spot being called upon to change abruptly between the two amplitude limits of black and white, it is required to alternate between these two levels in a sinusoidal manner. The visible effect of this limitation is that the pattern shows changes from black through all levels of grey up to peak white, and then down again, instead of reproducing the original sharp-edged black and white squares.

The British 625-line television system can do this satisfactorily if it is capable of handling sine wave modulation information up to the calculated frequency of 7·48 Mc/s.

However, in the statement of standards in Table 4.1, the highest modulation frequency for the British 625-line signal is seen to be 5·5 Mc/s. From this it is evident that the horizontal definition does not in fact equal the apparent value of the vertical definition. The reason for this discrepancy will now be dealt with.

**The Kell factor**

In practice the maximum vertical resolution obtainable is less than the number of active lines in the picture. This is confirmed by both subjective tests and by statistical analysis.

Subjective tests in this context are tests where observers look at reproduced television pictures and assess the quality (i.e. the sharpness and clarity) of vertical and horizontal details in the picture. Experience shows that observers assess the horizontal resolution as being as good as the observed vertical resolution, when the highest video frequency being used is in

fact *less* than the figure calculated by the methods described. If the bandwidth of the system is increased to accommodate this latter *calculated* figure, then the horizontal resolution is assessed as being *better* than the vertical resolution.

The reason for this is not hard to find. It can best be explained by reference to the chequerboard pattern of Fig. 4.7. To transmit this pattern the camera beam scans across it line by line. If it is the 'ultimate' pattern which is being scanned, the square elements are of line width. A perfect picture will be reproduced only if the scanning beam passes exactly along the centre of each horizontal line of squares (e.g. along PQ). If, however, the scanning beam should fall

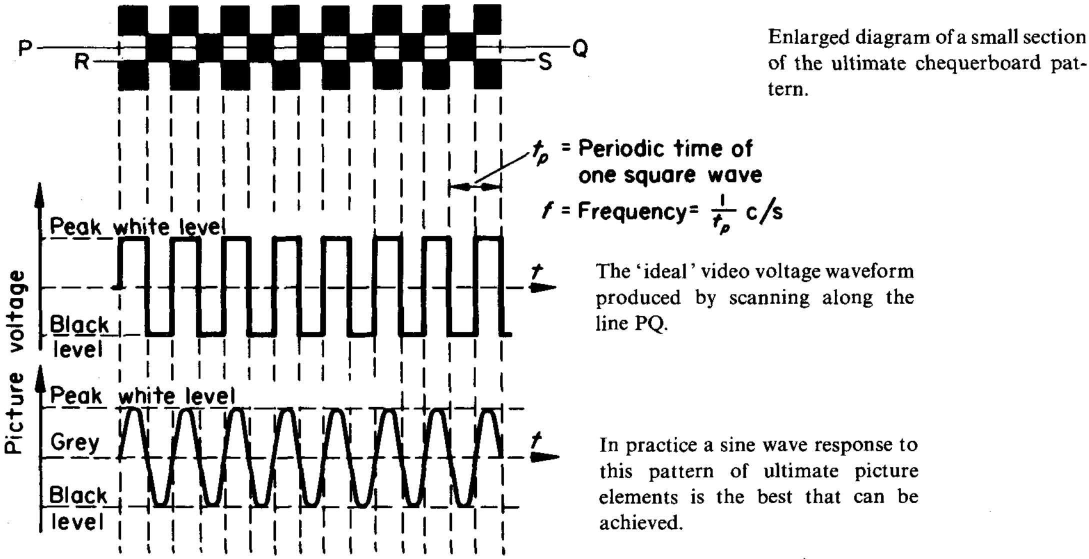



Fig. 4.7 Showing the 'ideal' and practical video voltage waveforms corresponding to the ultimate chequerboard pattern

along the dividing line between two rows of squares (e.g. along RS), then at any moment half of the scanning spot will be on white information and the other half on black. In this event the reproduced picture will be of a uniform grey, and no squares at all will be resolved. In an intermediate position, the beam will be more on one row than the adjacent one, and the squares will be discernible in the reproduced picture, though with diminished contrast.

Statistical analysis is said to suggest that the average number of effective lines is something of the order of 0·7 of the total active lines present. Subjective testing also appears to confirm that this is a reasonable assessment. This factor is known as the 'Kell factor'. It is obvious that it is not a precisely determinable quantity, and a variety of slightly different values are ascribed to it.

What it amounts to is that it is unrealistic to state that the vertical resolution is equal to the number of active lines, and then to proceed from that basis to calculate the required highest video frequency which must be accommodated to give an equal horizontal resolution. The first step is to recognise the fact that not all the lines (or all parts of an individual line) are fully effective *all* the time. The number of active lines must first be multiplied by the chosen Kell factor. This leads to a smaller figure for the inherent *actual* vertical resolution of the system.

The highest video frequency is then calculated to equate the horizontal resolution against this more realistic assessment of the available vertical resolution.

If this is done for the 625-line system discussed, a figure nearer the published standard is arrived at. Thus, suppose a Kell factor of 0·7 is chosen. The vertical resolution is then $0{\cdot}7 \times 585 = 409{\cdot}5$. A study of the working shows that the effect of the introduction of this factor is to make the final figure for the frequency equal $0{\cdot}7 \times 7{\cdot}5$ Mc/s $= 5{\cdot}25$ Mc/s. This compares with the actual standard of 5·5 Mc/s for this system.

The Kell factor is sometimes defined as being the ratio between the horizontal and the vertical resolution. This is a little misleading since it does not in itself give a clue to the true significance of the term. Thus to state that a given system has a Kell factor of 0·7 would appear from this definition, to suggest that the horizontal resolution is only 0·7 times as good as the vertical resolution. This of course is not true, since the whole argument has been directed towards making the practical horizontal and vertical resolutions equal.

The ratio would be better defined as the ratio between the horizontal resolution and the *maximum* possible vertical resolution.

To illustrate the point it is instructive to work backwards from the published standards of the British 625-line system, to deduce what apparent value for the Kell factor the standards imply. Thus:

$$\begin{aligned}
&\text{Maximum vertical resolution} = 585 \text{ elements}\\
&\text{Highest video frequency} = 5{\cdot}5 \text{ Mc/s}\\
&\text{Periodic time of one cycle} = \frac{1}{f} = \frac{1}{5{\cdot}5 \times 10^6} = \frac{10^{-6}}{5{\cdot}5} \text{ secs}\\
&\qquad = \frac{1}{5{\cdot}5}\,\mu\text{s}\\
&\text{Length of one active line} = 52\ \mu\text{s}\\
&\text{No. of complete cycles per active line} = \left(52 \div \frac{1}{5{\cdot}5}\right)\text{c/s}\\
&\qquad = (52 \times 5{\cdot}5) \text{ c/s}\\
&\qquad = 286 \text{ c/s}\\
&\text{No. of elements per line (i.e. no. of } \tfrac{1}{2}\text{-cycles)} = 2 \times 286\\
&\qquad = 572 \text{ elements}
\end{aligned}$$

This number of elements is accommodated along a line which is 4/3 times the picture height. To compare horizontal and vertical resolution, it is necessary to compare the number of elements contained in horizontal and vertical lines of equal length. Hence:

$$\text{No. of elements along } \textit{unit} \text{ horizontal length} = \frac{572}{4} = 143$$

$$\text{Also, maximum no. of elements along } \textit{unit} \text{ vertical length} = \frac{585}{3} = 195$$

$$\therefore \text{ Kell factor} = \frac{\text{Horizontal resolution}}{\text{Maximum vertical resolution}}$$

$$= \frac{143}{195} = 0{\cdot}73$$

It is interesting to note that if this procedure is repeated for the British 405-line system, for

which the published highest video frequency is 3 Mc/s, the Kell factor comes out to almost unity. This indicates a higher horizontal than vertical resolution with this signal.

When the number of lines is increased, however, the bandwidth automatically becomes greater. It then becomes more important to conserve bandwidth by keeping the video frequency as low as possible consistent with acceptable results. Under these circumstances, the introduction of a Kell factor then affords an economy in bandwidth.

The factors discussed in connection with the choice of the highest video frequency employed in a television system are recapitulated below.

**Picture definition**

1. The definition of a television system is a measure of its ability to give a clear, sharp reproduction of fine picture detail.
2. Definition depends upon both vertical and horizontal resolution.
3. Vertical resolution is a measure of the ability to reproduce detail in the vertical plane. The significance of this is seen if the detail along an imaginary vertical line through the picture, is considered. It is apparent that the maximum *vertical* resolution depends upon the number of *horizontal* lines present in the raster. This is true because the only way that detail along the imaginary vertical line can be resolved, is by changes in the brightness of the horizontal lines passing through the vertical line. The number of such details which can be resolved is obviously limited to the number of horizontal lines passing through the vertical line.
4. Horizontal resolution is a measure of the ability to reproduce changes in brightness *along* a horizontal line. Since such changes represent vertical edges of picture detail, it follows that horizontal resolution can be expressed as a measure of the ability to reproduce vertical information (or 'lines').
5. Whereas vertical resolution depends upon the number of raster lines employed in the system, horizontal resolution depends upon the rate at which the scanning spot is able to change its brightness level as it passes along a horizontal line. Horizontal resolution therefore depends upon the frequency bandwidth of the system; i.e. the maximum degree of horizontal resolution depends upon the highest video frequency accommodated by the system.
6. It is obviously logical to aim at equality of vertical and horizontal resolution. For this to be so the horizontal resolution must be such that the system is able to resolve a pattern of vertical lines of equal density to the horizontal lines of the raster. Since the picture is wider than it is high the number of vertical lines accommodated across one line of the picture, is equal to the number of active picture lines multiplied by the aspect ratio. For the purpose of the study, adjacent lines in both directions are assumed just to touch one another.
7. The maximum demand upon the system is made when alternate squares are black and white. When this is so the picture takes the form of a chequerboard pattern in which the individual squares have sides of length equal to the width of a raster line.
8. As the scanning spot moves along a line it is called upon to change in brightness level between black and peak white in a regular cyclic manner. Each pair of one black and one white square forms one 'square wave' cyclic change.
9. To change abruptly between black and white levels at this fast rate is demanding too much of the system and would necessitate a prohibitive bandwidth. It is technically and subjectively a fair compromise when dealing with very fine picture detail of this order,

to settle for a sine wave response to the 'square wave' pattern. Thus a fair approximation to the required pattern results if the scanning spot changes between the black and peak white levels sinusoidally as it traverses the screen, completing as many sine wave cycles during one line, as there are pairs of squares along the line.

10. To calculate the video frequency which corresponds with this sinusoidal change, it is first necessary to deduce the time duration of one cycle (i.e. its periodic time). This is done by dividing the known time duration of one active line by the number of cycles along the line. The reciprocal of the periodic time of the sine wave then gives the frequency in cycles per second.
11. The figure arrived at shows the highest video frequency which must be employed if the horizontal resolution is to be as good as the *maximum* possible vertical resolution. In practice, the actual vertical resolution experienced is less than the maximum value because only a proportion of the picture detail is precisely coincident with the scanning lines. From this it follows that the video frequency calculated, will give better horizontal resolution than the average vertical resolution.
12. To bring the horizontal resolution into line with a realistic assessment of the vertical resolution, the figure for the highest video frequency calculated by the method described, is multiplied by a Kell factor. A typical value chosen for this factor is 0·7. The Kell factor is determined on a statistical basis and its chosen value varies from system to system.

As a summary illustrating this method of arriving at an approximate idea of the order of the highest modulation frequency which ought to be accommodated for good quality image definition, the process is repeated below for the British 405-line signal.

| | |
|---|---|
| No. of lines per picture | = 405 |
| Field blanking period | = 14 lines |
| Time duration of active picture line | = (Line period—line blanking time) |
| | = $(98{\cdot}7-18)$ μs = 81 μs (approx.) |
| No. of suppressed lines per picture | = $(2\times 14) = 28$ |
| No. of active lines per picture (hence maximum vertical resolution) | = $(405-28) = 377$ |
| No. of vertical lines necessary for equal horizontal resolution | = $377\times$ aspect ratio |
| | = $377\times\frac{4}{3}$ |
| No. of cyclic changes along one picture line assuming alternate vertical lines are black and white | = $(377\times\frac{4}{3})\times\frac{1}{2}$ |
| Periodic time of one cyclic change | = $\dfrac{\text{Time duration of one active line}}{\text{No. of cyclic changes along one line}}$ |
| | = $\dfrac{81\times 10^{-6}}{377\times\frac{4}{3}\times\frac{1}{2}}$ seconds |
| | = $\dfrac{81\times 3\times 2}{377\times 4\times 10^{6}}$ seconds |
| Frequency of cyclic changes along a line | = $\dfrac{1}{\text{periodic time of one cyclic change}}$ |
| | = $\dfrac{377\times 4\times 10^{6}}{81\times 3\times 2}$ c/s(*) |
| | = 3·11 Mc/s |

This compares closely with the figure of 3 Mc/s quoted as the standard for the 405-line signal.

In this particular case the implicit Kell factor is approximately unity. From this it is evident that with the 405-line signal, the horizontal definition is better than the vertical.

With such a comparatively small number of lines, however, the necessary signal bandwidth is relatively small. There is therefore no pressing need to reduce the highest video frequency. With systems employing a greater number of lines, it becomes more important to restrict the highest video frequency to the lowest value consistent with observed equality of vertical and horizontal definition. For this purpose a Kell factor is introduced. An examination of the factors in the expression (*) above, shows that the following formula may be deduced:

Approximate highest modulation frequency for equal horizontal and vertical resolution

$$= \frac{\text{No. of active lines per picture} \times \text{aspect ratio}}{2 \times \text{time duration of one active line}}$$

When a Kell factor is introduced, it is evident that the above formula may be modified as follows:

$$\text{Highest video frequency} = \frac{\text{Kell factor} \times \text{no. of active lines} \times \text{aspect ratio}}{2 \times \text{time duration of one active line}}$$

**Influence of frequency response on sync. pulse shape**

Since synchronisation pulses are square waveforms of short duration, it follows that a wide frequency response is necessary if their shape is to be retained. As stated in the previous chapter, video signal standards include a specification of the maximum rise time for the sync. pulses. It is possible to estimate the frequency response necessary for the adequate reproduction of pulses, from a knowledge of the allowed maximum rise time.

These two factors are connected by the approximate formula:

$$\text{Highest necessary frequency} = \frac{1}{2 \times \text{allowed rise time}}$$

E.g. a typical figure specified for the rise time of television sync. pulses is 0·25 μs. To reproduce such pulses adequately, the circuitry handling the pulses should have a frequency bandwidth extending up to:

$$f = \frac{1}{2 \times 0{\cdot}25 \times 10^{-6}}\ \text{c/s} = \frac{10^6}{0{\cdot}5}\ \text{c/s} = 2\ \text{Mc/s}$$

It is clear that the pulses are safely preserved in video circuitry where, as has been shown, a frequency bandwidth considerably in excess of this figure has to be maintained in order to give the required picture definition.

**Influence of number of lines on bandwidth**

As the number of lines employed in a television picture is increased, the bandwidth necessary for a given quality of definition also increases. This is due to the fact that increasing the number of lines per picture (i.e. increasing the line frequency), decreases the time duration of each line. This means that the spot travels across the screen at a higher velocity, and the resolution of a given picture detail along a line involves a faster rate of change of the intensity of the scanning spot.

The matter is best demonstrated by taking a specific example.

In studying the 625-line system it was demonstrated that the resolution of 572 alternate black and white squares along a picture line involves the use of a video frequency of 5·5 Mc/s.

Suppose that we retain exactly the same line length and seek to resolve the same number of 572 picture elements, with a fictitious television system using twice as many lines (i.e. $2 \times 625 = 1250$ lines). Having doubled the line frequency, the line duration now becomes 32 μs instead of 64 μs. Suppose that the active line time is now 26 μs instead of 52 μs. The 572 square elements represented 286 complete cycles per line. The time duration of one cycle is now only 26/286 μs instead of 52/286 μs, i.e. one-half of what it was. Hence the frequency is doubled so that to set up an identical pattern requires the use of a video frequency of 11 Mc/s instead of 5·5 Mc/s.

This demonstrates a most important principle. It is seen that when the number of lines in a television picture is doubled, it is necessary to double the video frequency bandwidth in order to retain the *same* degree of horizontal definition as before.

Obviously the vertical definition increases as the number of picture line increases, but the horizontal definition gets worse unless the video frequency bandwidth is also increased.

In the case discussed it is worth pursuing the example a little further. Having doubled the number of picture lines, suppose that it is required to increase the horizontal resolution so that it again equals the vertical resolution. It was necessary to double the video bandwidth, merely to retain the *original* horizontal definition. To give the required equality of horizontal and vertical definition, since the number of vertical elements which can be resolved has been doubled, it follows that the system must now resolve $2 \times 572 = 1144$ elements along one line; i.e. 572 black-white cycles instead of 286. This clearly necessitates multiplying the original highest video frequency by a factor of four.

The result may be stated in general terms as follows: If the number of lines employed in a television system is increased, it is necessary to increase the video frequency bandwidth in direct proportion to the increase in the number of lines to maintain the *same* degree of horizontal definition as before. In order to increase the horizontal definition in the same proportion as the increase in the vertical resolution, the video frequency bandwidth must increase as the square of the increase in the number of lines.

**Comparison of systems**

It is instructive to compare the performance of various television systems to demonstrate the effect of different line standards.

The original British 405-line system is a convenient yardstick since it represents the smallest number of lines employed in any television system.

Here a video frequency bandwidth of 3 Mc/s is specified. Increasing the number of lines to 625 represents an increase of 1·54 times. To retain the *same* horizontal resolution as given by the 405-line system, the video frequency bandwidth needs to be increased to $1{\cdot}54 \times 3 = 4{\cdot}62$ Mc/s.

The European 625-line signal has a video bandwidth of 5 Mc/s which is only marginally more than the 4·6 Mc/s needed to give equality of horizontal definition with the 405-line signal. The British 625-line system has a video bandwidth of 5·5 Mc/s which is slightly better.

The French signal employs 819 lines, or almost exactly twice as many as the 405-line system. This system would require a video bandwidth of $2 \times 3 = 6$ Mc/s to give the *same* horizontal definition as the 405-line signal. In fact the bandwidth is 10·4 Mc/s so that this system has both a much improved vertical resolution and a better horizontal resolution.

The American 525-line system cannot be compared with such ease because the field frequency is also different. In this case it is necessary to compare the actual time durations of the active

picture lines. Table 4.1 shows the American line frequency to be 15,750 which compares very closely with the 625-line signal which has a line frequency of 15,625. From this it follows that the duration of one line of the 525 signal is very nearly the same as that of the 625 signal. Hence the bandwidth necessary to equal the 405-line horizontal definition is approximately the same as that deduced for the 625-line signal; i.e. 4·6 Mc/s. The standard actually employed is 4 Mc/s which shows that the horizontal resolution of this system is less than that of the 405-line system.

The comparison is interesting since it demonstrates that the number of lines employed by a given television system is not in itself, a guide to the quality of definition available from the system. It is true that the greater the number of lines the better the vertical resolution, but an assessment of the horizontal resolution afforded by a system requires a close scrutiny of the video bandwidth employed.

### Influence of picture frequency on bandwidth

The line frequency equals the product of the picture frequency and the number of lines employed in the system. Obviously then, if the picture frequency is increased whilst the number of lines per picture is held constant, the line frequency increases in proportion to the increase in picture frequency. It follows that to maintain the same horizontal resolution, the video frequency bandwidth must be increased in proportion to the increase in picture frequency.

In practice the picture frequency is almost universally constant at 25 c/s, to make the field frequency equal the mains frequency of 50 c/s. One exception is the American case where the picture frequency is 30 c/s to bring the field frequency in step with the 60 c/s mains. This system thus has less flicker than 50 c/s systems.

The influence of picture frequency on bandwidth was mentioned in Chapter 1. Here it was stated that flicker could be reduced by increasing the picture frequency, but to do this would also increase the bandwidth. This led to the introduction of the interlaced raster.

By splitting the picture into two fields, the beam is made to scan the picture vertically at twice the speed it would travel at with sequential scanning, but since the number of lines per second is the same as before, the bandwidth is unaltered. A further decrease in flicker could be achieved by a triple interlaced raster, but this would of course add to synchronisation problems.

# 5

# Vision Detectors

The general principles of television receivers and of television signals have been discussed. A convenient point in a receiver at which to start a detailed study of circuitry is the vision detector. It is at this stage that the video signal is extracted from the modulated R.F. vision signal. Previous to this stage is all the radio frequency circuitry. After it come the sections of the receiver which process the video signal, and obey its instructions.

Diode detectors are used almost exclusively for the demodulation of A.M. vision signals, as indeed they are in the great majority of normal A.M. broadcast receivers. A brief revision of the general basic pattern of these circuits is useful before looking at the specialised problems involved in the detection of vision signals.

The task of a diode detector is twofold:

(a) To rectify the modulated carrier wave.
(b) To filter out the R.F. (i.e. the I.F.) component of the rectified 'half' of the signal, and pass on a waveform corresponding to the shape of its envelope.

Three parts of a detector circuit may be discerned:

(1) The diode rectifier (either valve or semi-conductor).
(2) A *CR* time-constant network (usually called the reservoir capacitor and load resistor).
(3) An I.F. filter (which could be described as a 'low-pass filter').

The diagrams of Fig. 5.1 summarise the basic principles by building up step-by-step the normal form of a diode detector circuit. In Fig. 5.1(c) the effect of passing both unmodulated and modulated carrier waves to a simple diode and load resistor, is shown. Only the positive half-cycles of the carrier cause current through the diode so that the voltage across the resistor is a replica of just the positive half of the applied waveform. Since the output pulses which result from the A.M. wave vary in amplitude in sympathy with the envelope, it follows that the average output current varies at the modulating frequency. This simple circuit hence *does* 'detect', though very inefficiently. Its disadvantages are:

(1) The output contains a very strong I.F. component.
(2) The diode is fully conductive throughout the positive half cycles of *each cycle of the I.F. carrier*. This results in damping of the tuned circuit feeding the diode.
(3) The amplitude of the wanted modulation component is much less than the amplitude of the envelope (i.e. the modulation follows the average value of the I.F. pulses instead of following their peaks, as represented by the envelope).

The introduction of a 'reservoir' capacitor in parallel with the load resistor overcomes these disadvantages.

If the capacitor were present *without* the shunt resistor, then it would charge up after a few I.F. cycles to the peak value of the applied waveform, after which no further current would flow through the diode. With the load resistor in circuit, the capacitor discharges slightly through the resistor all the time. If the time constant of the *CR* network is long compared with the periodic time of the applied I.F. voltage, then the circuit settles down into a condition such that current only passes through the diode during the tip of each I.F. cycle to replace the loss of charge

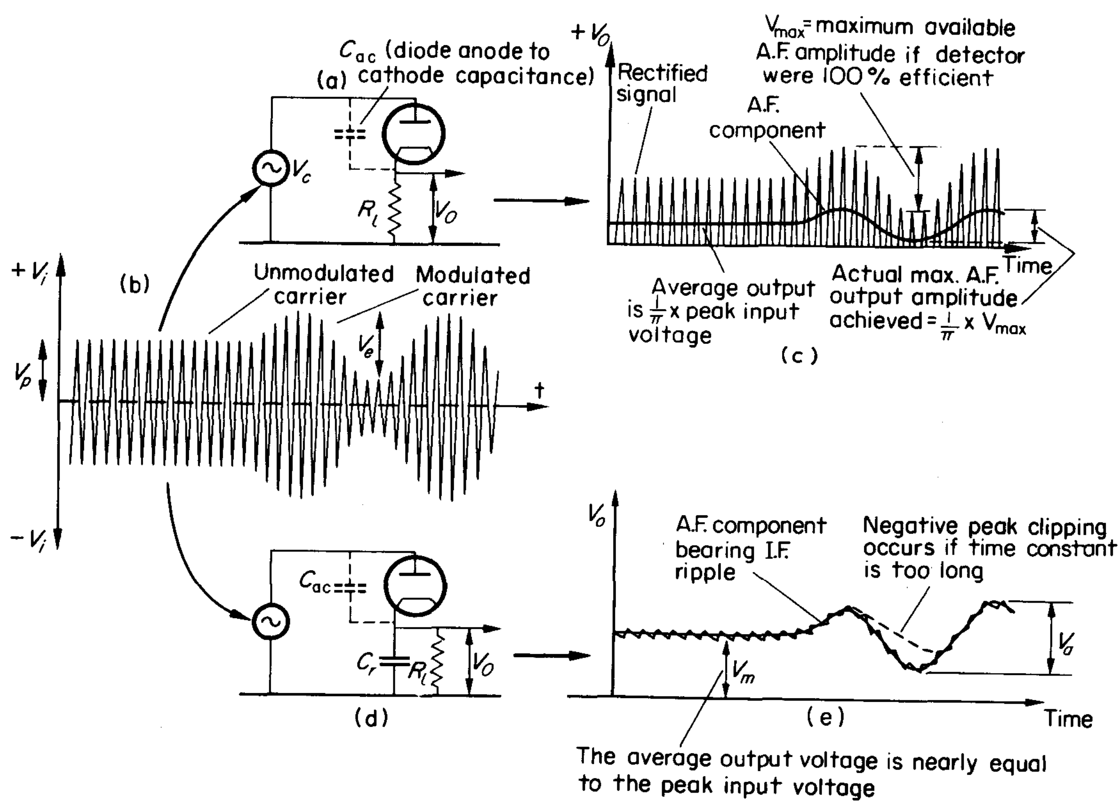



Fig. 5.1 Basic diode detector principles

*Note.* Efficiency $= \frac{V_m}{V_p} = \frac{V_a}{V_e}$ (see text).

which takes place through the resistor during the remainder of the cycle. If the applied waveform is an unmodulated carrier, then the voltage across the capacitor has an average value which is very nearly equal to the peak value of the applied voltage. This is shown in Fig. 5.1(e) from which it is seen that the 'output' voltage across $C_r$ is a 'direct' voltage with a superimposed I.F. ripple.

The time constant of the network must, however, allow the capacitor voltage to follow the comparatively slow 'undulations' of the envelope when the carrier wave is amplitude modulated. This is also shown in Fig. 5.1(e) where the modulation is assumed to be a simple audio frequency note.

The time-constant of the $CR$ network must, therefore, satisfy the following two requirements:

(a) It must be long compared with the periodic time of one cycle of the carrier wave.
(b) It must be short compared with the periodic time of the highest modulating frequency (i.e. it must be short enough to be able to follow the fastest rate of change of envelope amplitude).

*Example.* For broadcast radio, where an I.F. of the order of 450 kc/s is used, suitable values are $C_r = 100$ pF, $R_l = 500$ k$\Omega$.

That these values satisfy the above requirements may be demonstrated as follows:

Periodic time of one I.F. cycle is given by:

$$t = \frac{1}{f} = \frac{1}{450{,}000} \text{ seconds} = \frac{10^6}{45 \times 10^4}\ \mu\text{s} = 2{\cdot}2\ \mu\text{s}$$

Suppose a highest audio frequency of 10 kc/s is assumed. (In practice due to the restricted bandwidths on the congested broadcast bands, the figure is usually much less than this.)

Periodic time of one cycle of the 10 kc/s A.F. is:

$$t = \frac{1}{f} = \frac{10^6}{10^4}\ \mu\text{s} = 100\ \mu\text{s}$$

Time-constant of the $CR$ network is:

$$\begin{aligned} \text{t.c.} = CR \text{ seconds} &= 100 \times 10^{-12} \times 500 \times 10^3 \text{ seconds} \\ &= 5 \times 10^7 \times 10^{-12} \times 10^6\ \mu\text{s} \\ &= 50\ \mu\text{s} \end{aligned}$$

The time-constant is seen to be much longer than the I.F. periodic time, but shorter than the periodic time of the highest audio note. With a suitable time constant such as this, the audio output voltage obtained across the reservoir capacitor has an amplitude which is very nearly equal to that of the envelope of the applied A.M. wave.

A study of Fig. 5.1(e) shows that the output voltage consists of three components:

(1) The required audio frequency note.
(2) An I.F. ripple voltage superimposed upon the audio waveform.
(3) A d.c. component of amplitude almost equal to the average amplitude of the A.M. wave. (Note, by inspection of (b), that the average amplitude of the A.M. wave is equal to the *peak* amplitude of the unmodulated carrier.)

The I.F. carrier component is removed by passing the output signal through an I.F. filter. This is simply a low-pass filter which must be designed to offer low attenuation to all frequencies included in the modulation information but high attenuation to the I.F. component. For normal broadcast radio receivers this usually takes the form of a simple $CR$ filter. Fig. 5.2(b) shows a typical arrangement. $C_f$ and $R_f$ form a potential divider across the reservoir capacitor. The values of $C_f$ and $R_f$ are chosen so that:

(a) At the I.F. the reactance of $C_f$ is much less than the resistance of $R_f$.
(b) At A.F. the reactance of $C_f$ is much higher than the resistance of $R_f$.

When this is so the I.F. component is largely 'dropped' across $R_f$ and very little output at the I.F. appears across $C_f$; i.e. is passed on to the load resistor $R_l$. Conversely, the A.F. component is developed across $C_f$ and only a small proportion is 'dropped' across the series component $R_f$. Suitable values for the receiver described are $C_f = 100$ pF, and $R_f = 50$ k$\Omega$.

Finally the d.c. component is blocked in sound receivers by the insertion in series with the signal path of a d.c. blocking, A.F. coupling, capacitor. The d.c. component does, however, form a useful source of a.g.c. voltage since the amplitude of this component is a true indication of the strength of the incoming signal, and is independent of modulating information. Fig. 5.3 shows a typical diode detector circuit suitable for a normal broadcast receiver. To illustrate the effect of the various components, the form of the signal waveform is shown at three points in the circuit.

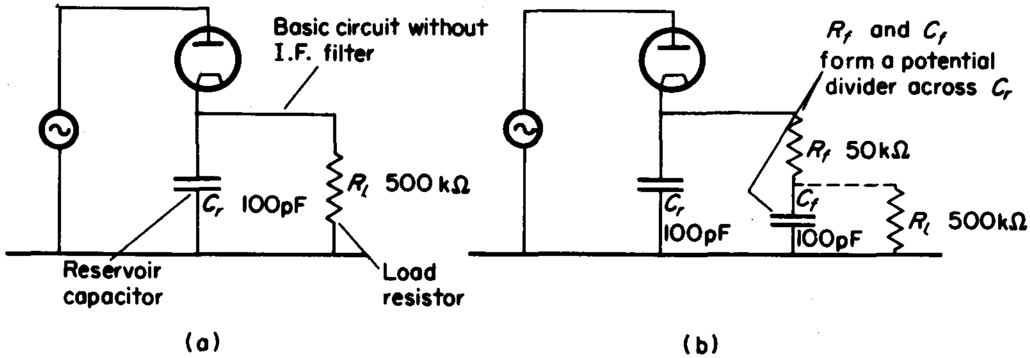



Fig. 5.2 Showing a simple resistance-capacitance I.F. filter used in detectors for broadcast radio receivers
*Note.* Reactance of $C_f$ at 450 kc/s = 3,537 Ω.
Reactance of $C_f$ at 5 kc/s = 318,400 Ω.

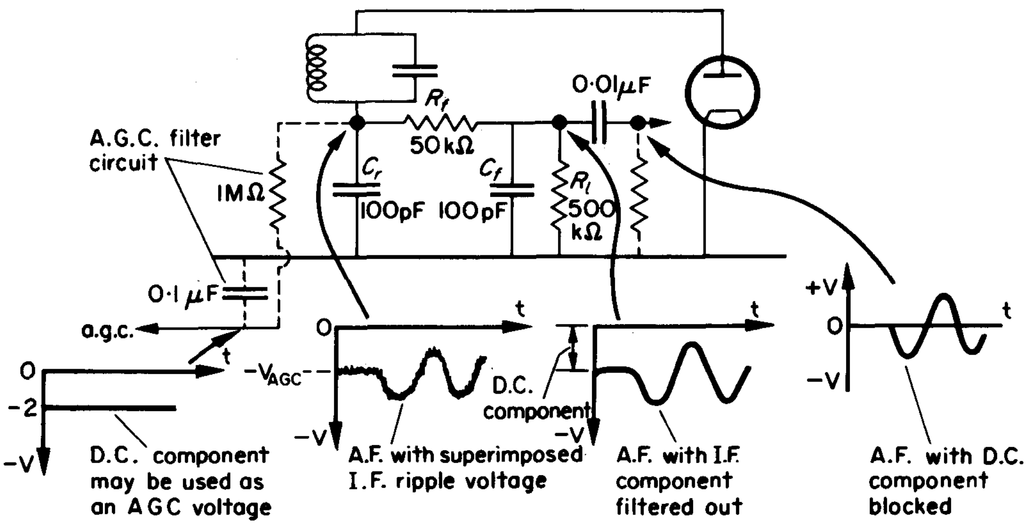



Fig. 5.3 Detector circuit suitable for broadcast receiver using I.F. of 450 kc/s
*Note.* Components are in the anode 'leg' to give a negative-going d.c. component for a.g.c. purposes.

**Efficiency**

The efficiency of a power-supply rectifier is simply expressed as follows:

$$\text{Efficiency of rectifier} = \frac{\text{Direct voltage output}}{\text{Peak alternating voltage input}}$$

This expression is equally applicable to a diode detector. Since, however, the object is to obtain an undistorted signal voltage output for a modulated carrier voltage input, the expression may perhaps with advantage be modified to the form:

$$\text{Detector efficiency} = \frac{\text{Peak-to-peak undistorted A.F. output voltage}}{\text{Peak-to-peak A.F. envelope input voltage}}$$

It should be noted that the denominator refers to the peak-to-peak amplitude of the A.F. envelope undulations of one half of the A.M. signal, and this dimension must not be confused with the overall peak-to-peak amplitude of the entire modulated wave.

A study of the waveforms in Fig. 5.1(b) and Fig. 5.1(e) shows that these two expressions are numerically equal. The first expression leads to Efficiency $= V_m/V_p$ and the second to Efficiency $= V_a/V_e$. This matter is referred to again in the comments which follow below upon negative-peak clipping.

In the ideal case, the diode would work as a switch which is *closed* and of *zero resistance* on positive half-cycles, but *open* and of *infinite resistance* on negative half-cycles. Under these conditions the efficiency of the simple circuit of Fig. 5.1(a) (i.e. without a reservoir capacitor) would be $1/\pi$. Thus the output is only approximately one-third of the total available amplitude. With a correctly chosen reservoir capacitor, the efficiency under these ideal conditions becomes almost unity (i.e. $\pi$ times as great).

In practice the forward resistance $R_d$ of the diode is not *zero* and nor, because of the presence of shunt capacitance, is the 'backward' impedance *infinite*. For sound radio, only the first of these factors is of significance. When the diode is conductive it is obvious that some part of the output voltage is dropped across its series resistance $R_d$. The efficiency then depends upon the ratio of $R_l/R_d$ where $R_l$ is the load resistance. (No I.F. filter network is assumed present in this discussion.)

For high efficiency the object is to make this ratio as high as possible. This involves:

(1) Choosing a diode with as low a forward resistance as possible.
(2) Making the load resistor $R_l$ as high as possible.

As $R_l$ is increased, $C_r$ has to be reduced to maintain the correct time constant. The smaller $C_r$ becomes, however, the more significant becomes the diode anode-to-cathode capacitance $C_{ac}$. Fig. 5.1(d) shows that $C_r$ and $C_{ac}$ form a potential divider across the input tuned circuit. On negative half-cycles the applied carrier voltage splits between these two capacitors.

It must be remembered that the voltage divides itself in inverse proportion to the capacitances, and the larger voltage appears across the smaller of the two series capacitors.

On that part of the cycle when the diode is cut off, therefore, instead of none of the applied voltage appearing across the load resistor $R_l$, the presence of $C_{ac}$ results in part of the unwanted 'negative half' of the applied voltage being developed across the reservoir capacitor. This reduces the net positive-going output and hence the efficiency. When $C_{ac}$ is very much less than $C_r$, the relatively very high reactance of $C_{ac}$ and low reactance of $C_r$ results in virtually all the applied voltage being 'dropped' across the diode and none across the output circuit (i.e. the diode approximates towards the ideal 'open-circuit' condition).

This argument, though over-simplified, serves to show that it is desirable to maintain $C_r$ to be substantially greater than $C_{ac}$ in order to preserve efficient operation. This in turn points to one reason why $R_l$ may not be increased unduly.

In practice, a value for $C_r$ of the order 50 to 100 pF is common in normal valve A.M. broadcast receivers.

**Negative-peak clipping**

Before leaving the question of efficiency it is important to keep in mind that the object of a detector is to produce an undistorted output at the envelope frequency. If $R_l$ were to be increased, without modifying the value of $C_r$, then the time constant would become too long to allow the network to follow the envelope changes at high modulation frequencies. It should be noted that $C_r$ charges through the diode but discharges through $R_l$. An excessive time constant does not therefore affect the positive-going half-cycles of the envelope waveform (since $C_r$ follows these by charging through $R_d$), but does cause distortion of the negative-going movements. The effect is shown in Fig. 5.1(e) where the dotted line indicates how the voltage across $C_r$ would fail to follow the envelope shape if the time constant did not meet the requirement of being shorter than the periodic time of the highest modulating frequency. The effect is known as negative-peak clipping.

The first expression for efficiency of rectification quoted above relates the d.c. output to the peak a.c. input. It is evident that increasing $R_l$ without increasing $C_r$ must increase the d.c. output and hence the efficiency.

In using this ratio to determine *detector* as distinct from *rectifier* efficiency, however, the need to specify an undistorted output must not be forgotten. For this reason, the second expression, which relates the amplitude of the undistorted A.F. output to the amplitude of the A.F. 'envelope' input, is more meaningful.

**The diode vision detector**

Having looked at general principles it is easy to progress to study the modifications necessary to make the circuit suitable for televison receivers.

The factors which now influence the design are:

(1) The I.F. is very much higher (e.g. of the order 30 to 40 Mc/s instead of 450 kc/s).
(2) The bandwidth of the video modulation is very much greater (e.g. extending up to several Mc/s instead of say 10 kc/s).
(3) The ratio of the I.F. to the highest modulation frequency, is much less (e.g. a ratio of 40 Mc/s:5 Mc/s=8:1, instead of 450 kc/s to 10 kc/s=45:1).

As a basis for discussion, suppose that the vision I.F. is 35 Mc/s and the highest video frequency is 5 Mc/s.

*Choice of $R_l$ and $C_r$*

The periodic time of one I.F. cycle is now:

$$t = \frac{1}{35\times 10^6}\text{ seconds} = \frac{1}{35}\ \mu\text{s} = 0{\cdot}03\ \mu\text{s}$$

The periodic time of one cycle of the highest modulation frequency is:

$$t = \frac{1}{5\times 10^6}\text{ seconds} = \frac{1}{5}\ \mu\text{s} = 0{\cdot}2\ \mu\text{s}$$

The time constant of $C_r$ and $R_l$ should therefore be greater than 0·03 μs but smaller than 0·2 μs.

The corresponding figures for the broadcast radio receiver were 2·2 μs and 100 μs. In the latter case a time constant of 50 μs (or half the periodic time of the highest modulating frequency) was chosen. It is clear that component values for $C_r$ and $R_l$ have to be much different for the vision signal. Suppose a time constant of 0·1 μs is aimed at. (As before this is half the periodic time of the highest modulating frequency.) It now becomes necessary to decide upon values for $C_r$ and $R_l$.

The following relevant factors must be remembered:

(1) For maximum efficiency the ratio of the load resistor $R_l$ to the diode's forward resistance $R_d$ must be as large as possible.
(2) When $R_l$ is increased, however, $C_r$ must be decreased to maintain the required time constant.
(3) But the ratio of $C_r$ to the diode's capacitance $C_{ac}$ must be as high as possible.

A compromise is needed between (3) and (2). Suppose a value for $C_r$ of 20 pF is aimed at. Part of this will be stray capacitance so that the 'fitted' capacitor would show a smaller value (e.g. 10 pF or less). Hence the required value for $R_l$ is:

$$R_l\ (\text{ohms}) = \frac{\text{Time constant in seconds}}{C_r\ \text{farads}}$$
$$= \frac{0{\cdot}1 \times 10^{-6}}{20 \times 10^{-12}}\,\Omega = \frac{10^4}{2}\,\Omega = 5000\ \text{ohms}$$

Values similar to these will be found in practical circuits. It is of course difficult to assess how much of the effective value of $C_r$ is due to stray capacitance in any given circuit but examples of *fitted* values for $C_r$ and $R_l$ are shown in Table 5.1.

**Table 5.1**
*Examples of values for $C_r$ and $R_l$ in practical vision detectors*

| Reservoir capacitor $C_r$ | Load resistor $R_l$ | Vision I.F. | Highest video frequency |
|---|---|---|---|
| 8·2 pF | 4·7 kΩ | 39·5 Mc/s | 5·5 Mc/s |
| 5 pF | 3·9 kΩ | 39·5 Mc/s | 5·5 Mc/s |
| 5 pF | 2·2 kΩ | 39·5 Mc/s | 5·5 Mc/s |
| 4 pF | 2·7 kΩ | 38·9 Mc/s | 5 Mc/s |
| 8·2 pF | 5·6 kΩ | 34·65 Mc/s | 3 Mc/s |
| 4·7 pF | 6·8 kΩ | 34·65 Mc/s | 3 Mc/s |

The values of $C_r$ and $R_l$ have been studied on the basis of time constants. That $R_l$ must be comparatively low is also obvious when impedances are considered.

For good frequency response it is necessary that the impedance of any network handling the video signal should not vary widely over the frequency range contained by the signal. If

$R_l$ were of high value the effect of shunt capacitance would be to cause an increasing fall-off of output at upper frequencies. Since the video signal contains frequencies up to several Mc/s, and shunt capacitance across the signal path is inevitable, the only way in which the high frequency response may be preserved is to use low values of resistance so that the net impedance remains more nearly constant.

Sometimes a small inductor in series with $R_l$ gives a lift to the impedance of this arm of the circuit at the high frequency end of the range, so partly compensating for the increased shunting effect of the capacitance.

*Choice of diode*

It is evident from the values of $C_r$ and $R_l$ deduced above, that the following qualities must be sought in choosing the diode:

(1) Forward resistance must be as low as possible since $R_l$ is only a few thousand ohms.
(2) $C_{ac}$ must be as low as possible.

The inter-electrode capacitance of normal valve diodes used at broadcast frequencies is much too high. Special low capacitance diodes are available which may be used as vision detectors (e.g. EB91). With such valves $C_{ac}$ is of the order of 3 pF. However, semi-conductor diodes (e.g. OA70) are widely used and with these $C_{ac}$ is reduced to the order of 1 pF. Typical parameters for a semi-conductor diode suitable for this purpose are:

| | |
|---|---|
| $C_{ac}$ (at zero applied voltage) | = 1 pF |
| $R_d$ | = 400 Ω |
| Maximum peak Inverse voltage | = 22·5 V |

Semi-conductor diodes have other advantages. They are very small and require no heater connections. It is therefore possible to enclose the diode detector circuit in the last vision I.F. transformer screening can. A 'non-linear' impedance, such as a diode, produces a whole series of harmonic and beat frequencies when fed with R.F. signals. The vision detector receives an I.F. input of several volts and many unwanted frequency components are present at significant strength in the output of the detector. It is possible for these unwanted products of the detection process to radiate back to the tuner unit to give rise to annoying interference. For example, a harmonic of the I.F. may fall near enough to the carrier frequency of the received signal to give rise to an output component from the frequency changer, which is not far removed from the original vision I.F. The beat frequency produced when this and the true I.F. reach the detector may fall within the video signal bandwidth and go forward to the c.r.t. to cause patterning on the picture. By containing the detector circuit in a screening can, and also by the use of suitable I.F. filter circuitry, this danger is obviated.

One disadvantage of semi-conductor diodes compared with valves is their susceptibility to damage by excessive inverse voltages; e.g. in the example quoted, the peak inverse voltage must not exceed 22·5 volts. The low-forward-resistance, low-capacitance diodes suitable for use as v.h.f. detectors, tend to have comparatively low peak inverse voltages of this order.

**I.F. filters**

The resistance-capacitance filter circuits used at broadcast frequencies are replaced by inductance-capacitance filters in vision detectors. This is necessary because of the low value of load resistor used in the vision detector. To place a series filter resistor in the circuit is clearly impracticable, since if this had a value high enough to attenuate the I.F., it would seriously

attenuate the video signal as well. A series inductor is therefore inserted in place of the resistor. Fig. 5.4(b) shows a basic circuit. Here the filter elements are $L_f$ and $C_f$ and the values of these components are chosen so that:

(a) At the I.F. the inductor $L_f$ has a much higher reactance than the shunt capacitor $C_f$.
(b) At video signal frequencies the reactance of $L_f$ is much lower than that of $C_f$.

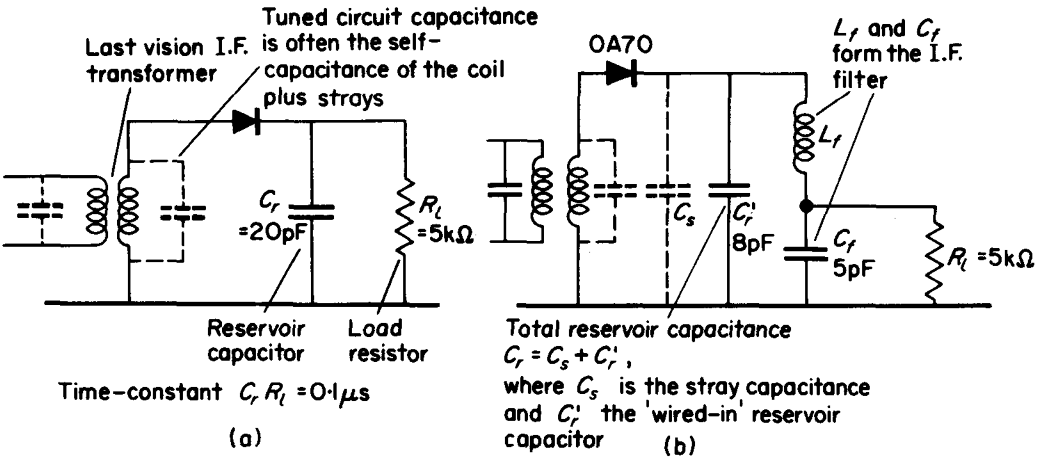



Fig. 5.4 Vision detector circuit development
(a) Shows a basic circuit with typical values for $C_r$ and $R_l$.
(b) Shows the inclusion of an I.F. filter.

As with the broadcast receiver R.C. filter, this results in the I.F. being 'dropped' across the series element of the filter but the modulation component is passed on to the load resistor.

Various subtleties are possible with this simple circuit. For example,

(1) $L_f$ may be of such a value that it resonates with its own shunt capacitance $C_s$ at the I.F. Since the impedance of a parallel tuned circuit is maximum at resonance, the circuit forms a high series impedance to the I.F.
(2) $L_f$ may be made to resonate with the filter capacitance $C_f$ at a frequency just above the highest frequency contained in the video signal. The current through a series tuned circuit is maximum at resonance and the voltage across $C_f$ rises to a maximum. The circuit is of low $Q$ but the device does give some degree of 'lift' to the response curve at the upper end and may be used to correct a response which otherwise falls off in this region.
(3) In some cases the required filter capacitance is provided by the input capacitance to the video amplifier stage so that no 'wired-in' $C_f$ appears in the circuit.

Often, as illustrated in Fig. 5.5, more than one series inductor is used, to give more efficient filtering. Again it is possible to make the inductors behave as tuned circuits. As an example, $L_{fa}C_a$ may resonate at the I.F. whilst $L_{fb}C_b$ may be made to resonate at a particularly troublesome harmonic of the I.F. Whether or not the inductors *do* form such deliberately designed tuned circuits cannot of course be seen from a circuit diagram, nor is it possible to treat these circuits as sophisticated 'T'- or '$\pi$'-section filters which are designed according to rigorous filter

theory. In essence they are simply an arrangement of series and shunt impedances which offer a satisfactory measure of attenuation to unwanted frequencies but a not-too-serious attenuation of the video signal frequencies.

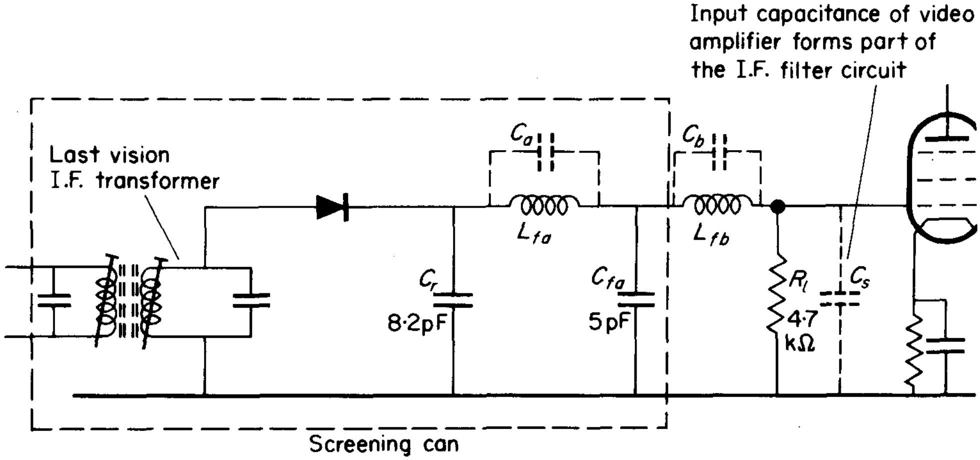



Fig. 5.5 Vision detector circuit with a two-section I.F. filter

**Polarity of the output signal**

This subject has already been introduced in Chapter 2. Before pursuing the matter further it is very important to revise this introductory material. The text concerned is under the section headed 'Negative Modulation' and this together with the diagrams of Fig. 2.6 should be studied carefully before proceeding. (See pages 24, 25 and 26.)

The factors which determine what polarity the video signal output from the detector must have are:

(a) The number of video amplifier stages between the detector and the c.r.t.
(b) The c.r.t. electrode to which the video signal is fed (i.e. the grid or the cathode).

If, as is usually the case, the video signal is fed to the c.r.t. cathode, then a negative-going video signal is required at the tube. In this case, as the video voltage level moves from black-level towards peak-white, the tube cathode is driven less positive to the grid, and the beam current increases. Conversely, if grid modulation is used, a positive-going video signal must reach the tube.

Since each amplifier stage inverts the signal, it follows that when an *even* number of stages is used, the input to the first stage must have the same polarity as that required by the c.r.t. With an *odd* number of stages the detector must deliver to the first video amplifier a signal of opposite polarity to that required by the tube. (It should be remembered that cathode-follower and emitter-follower stages do *not* invert the signal, and are therefore not counted in this context.)

In a large number of valve receivers, a single video amplifier stage feeds a negative-going signal to the cathode of the c.r.t. A positive-going video signal must be delivered by the detector to the grid of the video amplifier in such cases.

In Chapter 2, Fig. 2.6 shows how positive-going signals may be obtained from either positively or negatively modulated vision signals.* The fact that the positive-going video signal

* See Fig. 2.6(a) and Fig. 2.6(j) respectively.

is entirely positive to chassis potential in the one case, but negative to it in the other, leads to necessary differences in the biasing arrangements of the video amplifier. These are now dealt with.

*Arrangements for positively-modulated signals*

Under 'no-signal' conditions, the video amplifier is biased back towards the foot of the $I_a/V_a$ curve, as shown in Fig. 5.6. The video signal is taken from the cathode of the diode and d.c. coupled to the control grid. Under these circumstances the positive-going video signal

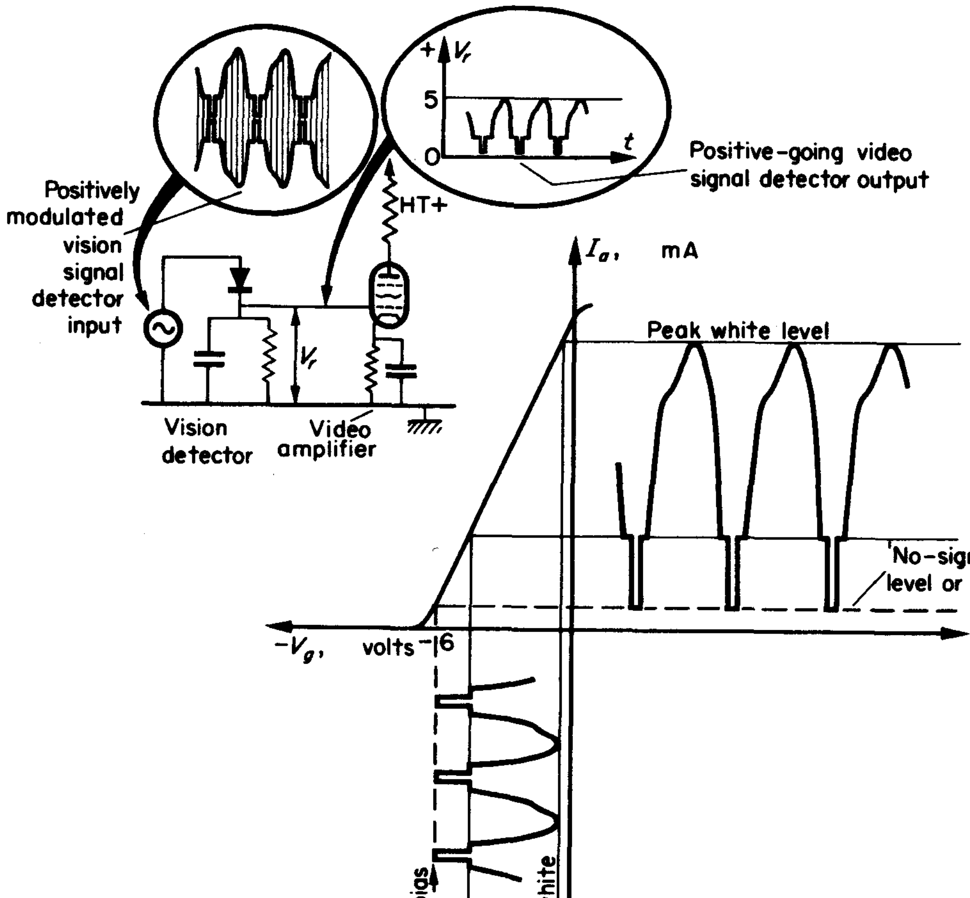



Fig. 5.6 Showing the working conditions of a video amplifier when handling a d.c. coupled positive-going video signal derived from a positively modulated carrier. Note the low level of 'no-signal' anode current

lessens the net bias on the video amplifier and causes it to conduct harder. The actual net grid-cathode voltage at any instant is the algebraic sum of the cathode-to-chassis voltage and the instantaneous video signal grid-to-chassis voltage. For example, suppose the no-signal

cathode bias voltage is 6 volts. This causes the grid to be 6 volts negative to the cathode. If a peak-white video output signal is now developed, such that the voltage presented by the detector to the amplifier grid is 5 volts positive to chassis, then the net bias on the valve during peak white information is $-6+5=-1$ volt.

It should be noted once again that the video signal obtained at the cathode of the diode is both positive-going in the sense that increasing whiteness drives the voltage output positive with respect to its value at black-level, and also positive-going in the sense that all parts of it are positive to chassis.

The circuit of Fig. 5.5 is typical of those found in positive modulation receivers (e.g. British 405) in which the video signal is fed via one video amplifying stage to the c.r.t. cathode.

*Arrangements for negatively modulated signals*

In contrast to the previous case, to achieve a positive-going video signal, the output must be

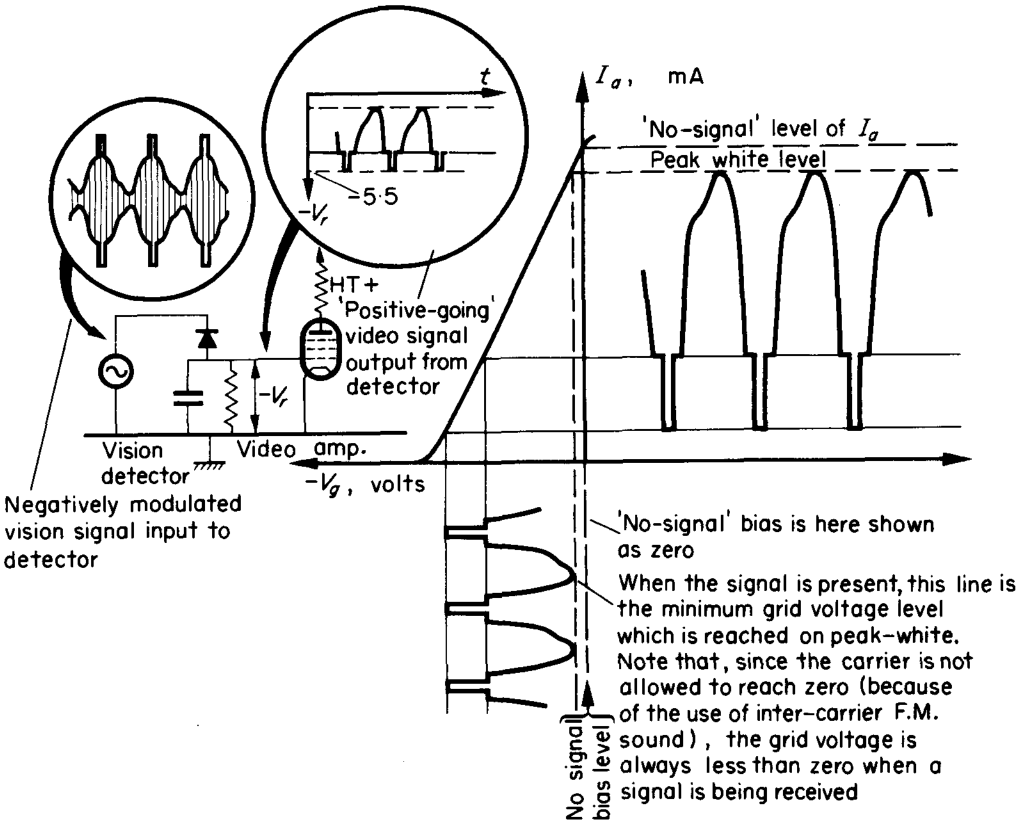



Fig. 5.7 Showing the working conditions of a video amplifier when handling a d.c. coupled positive-going video signal derived from a negatively-modulated carrier

To clarify the matter, the no-signal bias voltage is shown as zero. In practice a small fixed bias usually moves the no-signal working point just back from zero.

Compare the high level of video amplifier's no-signal anode current with the low level shown for the positively-modulated case in Fig. 5.6.

taken from the diode anode. This was shown in Fig. 2.6. The very important difference in the nature of the output signal must be clearly understood. Thus, whilst the signal arrived at is of the required positive-going form in the sense that the peak-white voltage is positive to that at black-level, it is none the less entirely *negative* with respect to chassis potential.

If d.c. coupling is again used between the detector and the video amplifier control grid, the no-signal bias must be either zero or very low, whereas in the positive modulation receiver the valve was biased back nearly to cut-off. This is shown in Fig. 5.7 in which, to make the point clear, zero fixed bias is assumed.

When a signal is being received it now drives the control grid negative to earth, hence increasing the net bias on the valve. The maximum value of this negative drive voltage occurs during sync. pulses when the carrier is 100% modulated. At these times the net bias on the video amplifier approaches cut-off. As picture detail moves from black-level to peak-white level, the negative drive voltage on the grid diminishes, and the video amplifier anode current increases.

A little thought shows that it is only the datum line which has changed. The video signal fits beneath the $I_a/V_a$ curve as before, and a move from black-level to white-level produces precisely the same effect in the two systems.

It is in the conditions which exist during the absence of a signal that the major difference occurs.

With a positive modulation receiver, in the absence of a signal, the video amplifier valve is biased right back and passes only a low current, so that the anode voltage remains high. Assuming d.c. coupling to the c.r.t. cathode, the latter is then at its maximum positive voltage relative to the tube grid, and only a low beam current flows.

Conversely, in a negative modulation receiver, when no signal is being received, the video amplifier valve has little or no bias. This leads to maximum anode current and minimum anode voltage. The c.r.t. cathode is at its *minimum* positive voltage and the beam current is at maximum. It is obviously not desirable to have the tube running at maximum brightness at these times. In order to overcome this disadvantage, a.c. coupling is sometimes used between the detector and the video amplifier in negative modulation receivers. This has the effect of removing the d.c. component of the video signal.

Before passing on to discuss a.c. coupling, it is instructive to study examples of direct-coupled detectors for negative modulation receivers.

In Fig. 5.8 a detector suitable for the British 625-line signal is shown. The reservoir capacitor $C_r$ is of 8 pF, and the load resistor is 3·3 kΩ. In series with the load resistor is a compensating coil $L_c$ which improves the high frequency response. $L_a$, $C_a$ and $L_b$ with the stray shunt capacitance $C_b$ form the two-section I.F. filter.

The detector circuit is also seen to include a 6 Mc/s intercarrier sound I.F. trap-circuit. The basic principle of the intercarrier sound I.F. system was explained in Chapter 2. The inter-I.F. beat frequency component produced at the detector is sometimes extracted directly from the detector circuit, as in this case, but is often allowed to pass through the video amplifier to receive this additional amplification before being passed to the F.M. sound receiver I.F. stage(s). The method of extraction is simply to insert a parallel tuned circuit at a suitable point somewhere along the signal path. Such a circuit is tuned to the inter-carrier difference frequency and offers a high impedance to this component so that it is effectively removed from the video signal path. A tuned secondary circuit passes on the intercarrier I.F. to the sound receiver circuits. The inter-carrier frequency is of course equal to the frequency which separates the transmitted vision and sound carriers. This is 6 Mc/s with the British 625-line system;

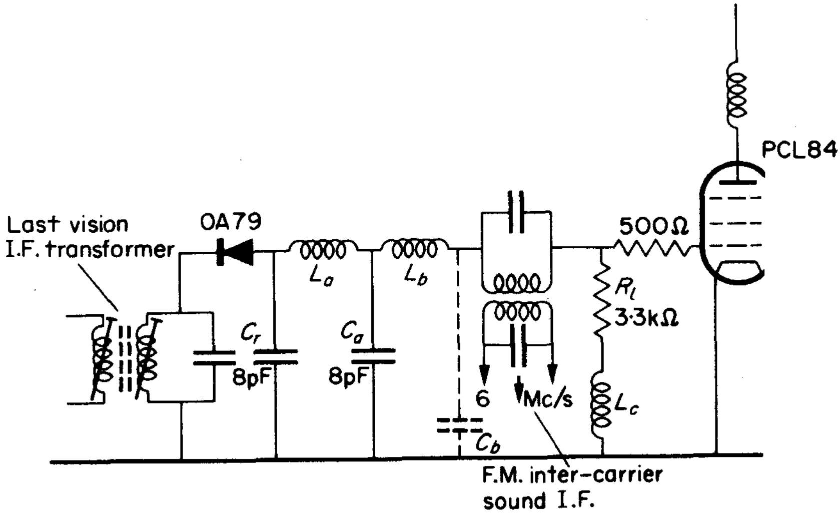



Fig. 5.8 Example of a d.c.-coupled detector suitable for a British 625-line negative-modulation receiver

5·5 Mc/s with the European 625-line system, and 4·5 Mc/s in the case of the American 525-line system.

A similar circuit from an American 525-line receiver is illustrated in Fig. 5.9. Here the reservoir capacitor is of 10 pF and the 2·2 kΩ load resistor is in series with a 125 μH compensating coil. Three series inductors are used in the filter circuit; two of them being tuned by their own self-capacitance to resonate at I.F. harmonics. In this case the 4·5 Mc/s inter-carrier I.F.-extraction tuned circuit is connected, via a 2·2 pF capacitor, in shunt with the detector, rather than in series with it.

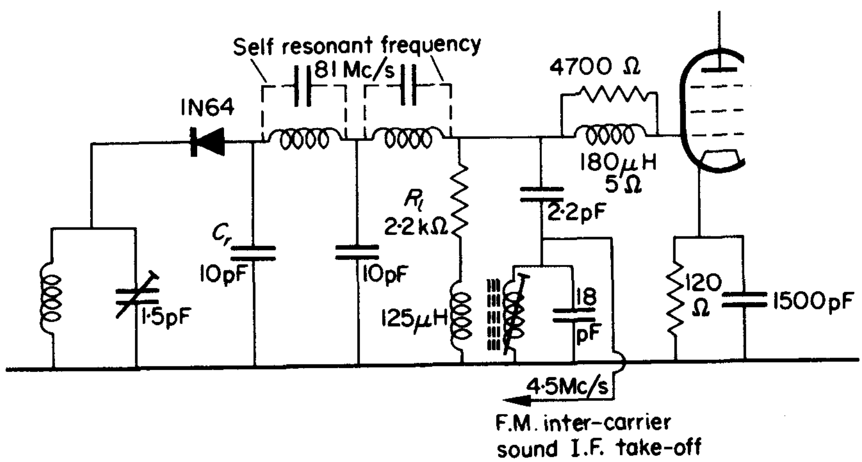



Fig. 5.9 Example of a d.c.-coupled detector in an American (525-line) negative-modulation receiver

### The d.c. component

Unlike the A.F. modulation in a sound receiver, a video signal includes a d.c. component. For a *true* reproduction of the transmitted picture, this d.c. component must be preserved. This is illustrated in Fig. 5.10, which shows two different lines of a given picture; one assumed grey, and the other white. The voltage corresponding to black-level is the same for both lines. This constant black-level voltage represents the reference level against which all picture detail is measured.

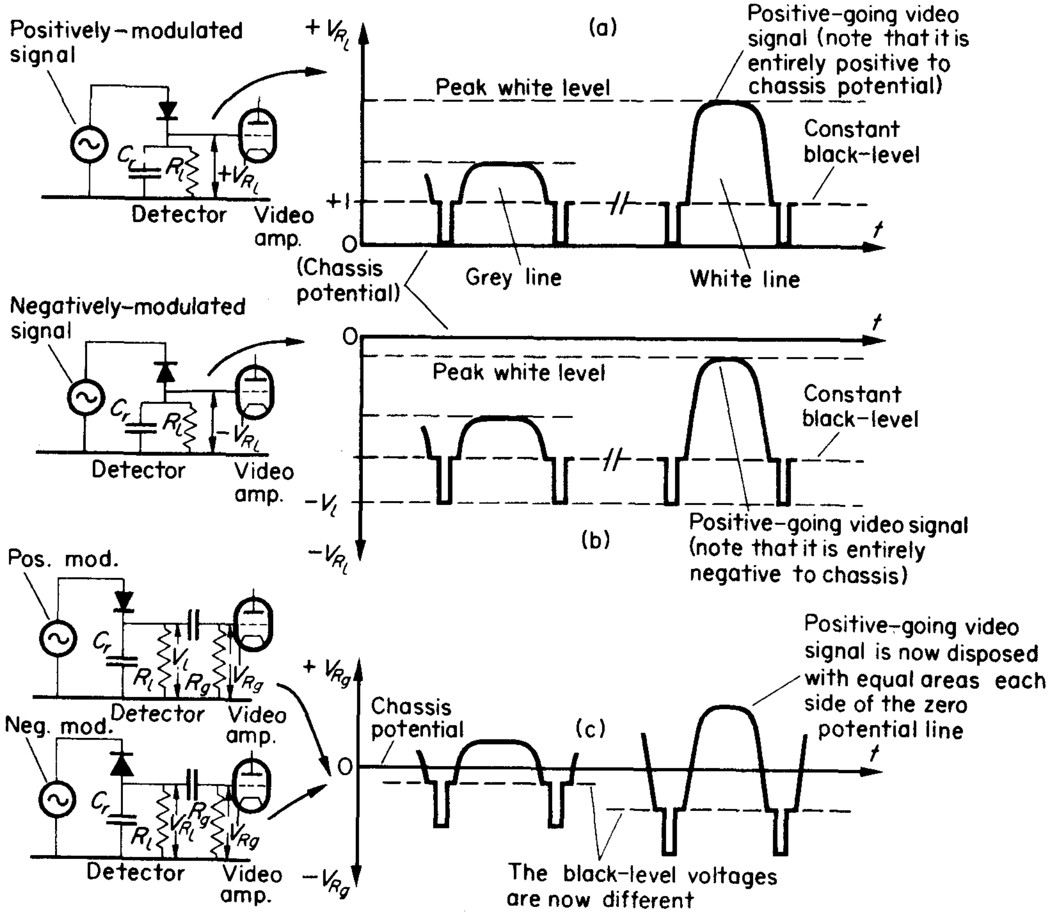



Fig. 5.10

(a) Positively-modulated case

The detector output voltage for two different lines, one grey and the other white, is shown. Black-level is seen to be a constant voltage level which is not affected by picture detail.

(b) Negatively-modulated case

The detector output for the same two lines is shown. Again, black-level is a constant voltage level.

(c) Effect of a.c. coupling

The video waveforms settle with equal areas on each side of the zero potential line, and the voltage corresponding to black-level changes with picture detail. Thus the d.c. component is lost. Note that the final waveforms are identical from both systems. This is to be expected, since the *only difference* between (a) and (b) is that the d.c. components are positive and negative to chassis respectively, and a.c. coupling removes these d.c. components.

Provided the signal strength does not change, black-level remains constant and both picture detail on one side, and the sync. pulses on the other, may be regarded as voltage excursions which are measured with respect to this steady 'd.c.' reference line.

In Fig. 5.10(a) the detector output from a positively modulated carrier is shown. Since the tips of sync. pulses are at zero carrier level, the video signal output waveform 'sits' upon the zero voltage line. Black-level is shown as +1 volt in this diagram. The fact that the two video signal lines are both sitting on the same voltage level base (in this case 0 V) allows the difference between the picture detail of the two lines to be clearly seen as a voltage-level difference. If the detector is d.c. coupled to the grid of the video amplifier, this true comparison is preserved since the anode current waveforms again 'sit' upon a common base line formed by the no-signal value of $I_a$.

Fig. 5.10(b) shows the same two video signal lines as they would appear in the output of the detector in a negative modulation receiver. Since the carrier is always at its maximum 100% amplitude during sync. pulse tips, it follows that the negative voltage developed across the detector load resistor is at its maximum negative value during sync. pulses. The video signal thus sits upon this particular negative voltage line, and black-level is once again represented by a constant voltage as shown. Increasing degrees of whiteness correspond to decreasing values of negative voltage across the resistor; i.e. the video signal is positive-going. As before, the grey and white lines are correctly reproduced by the receiver, if d.c. coupling is employed from the detector to the video amplifier and from the video amplifier to the tube. Assuming the video amplifier is correctly biased in each case, the positive and negative modulation receivers produce identical video signal outputs at the amplifier anode.

The significance of the d.c. component is further discussed in subsequent chapters, but sufficient has been said to allow a return to the main theme of this chapter.

**A.C. coupling from the detector to the video amplifier**

A.C. coupling is sometimes used in negative modulation receivers to overcome the disadvantage of high beam current under no-signal conditions.

The insertion of a coupling capacitor in series with the signal path to the video amplifier grid completely removes the d.c. component. This is illustrated in Fig. 5.10(c), and it is interesting to note that the signals produced across the grid resistor by the outputs from the positive modulation and negative modulation detectors are identical.

The video signal waveform settles down with equal areas either side of the zero-voltage line. The grey and white lines now appear at the video amplifier grid, in the form shown on the diagram. With the white line, in order to dispose the waveform equally about the zero voltage line, black-level must descend to a more negative potential than that of the grey line. The result is that the contrast between the two lines is less than it should be. An increase in the average brightness of the transmitted scene results at the receiver in a depression of black-level, so that the reproduced change of brightness is less than that of the original scene.

To accommodate an a.c.-coupled input, the video amplifier needs to be biased to the mid-point of the linear part of its dynamic $I_a/V_g$ curve. This represents one advantage of using this form of coupling in a negative modulation receiver, since the no-signal anode current of the amplifier is much less than it has to be in the d.c. coupled case. This results in the anode voltage, and with it the c.r.t. cathode voltage, being more positive under no-signal conditions, and hence the beam current is much less.

Fig. 5.11 shows the video amplifier operating conditions. The anode current corresponding to black-level is different for the two lines and must clearly vary as the mean level brightness

of groups of lines varies. Suppose, for example, that the bias on the video amplifier is fixed to give correct reproduction of a very bright scene. This would correspond to the way in which the peak-white line is positioned on the diagram. If this scene is now followed by a darker one, black-level moves in under the $I_a/V_g$ curve, and the picture is relatively brighter than it should be. Of course, black-level cannot 'jump' instantaneously from one voltage level to another, because of the time constant of the a.c. coupling network. If a large number of white lines are transmitted, the working condition will settle down as at A in Fig. 5.11. If these are now

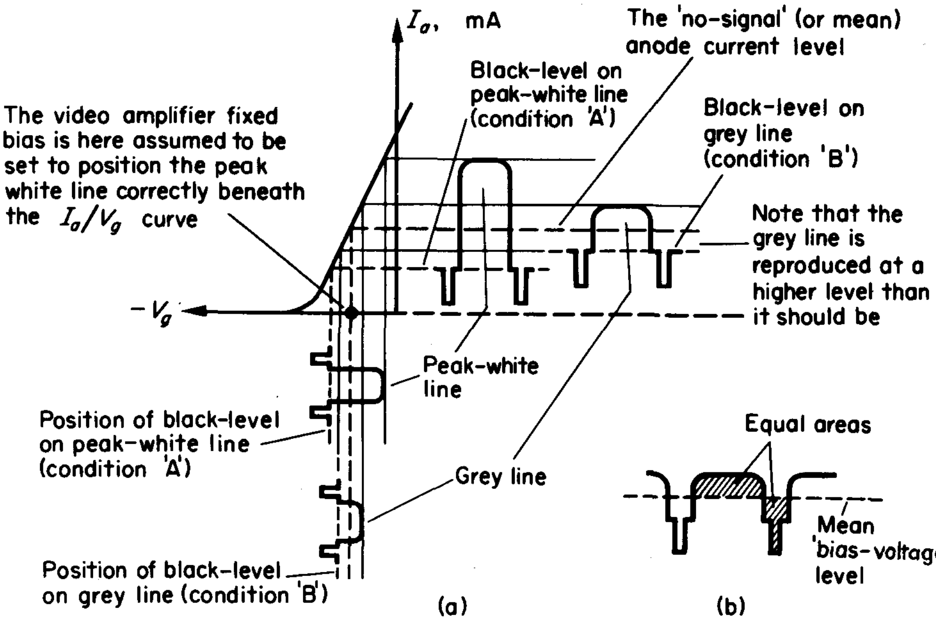



Fig. 5.11 Illustrating how it is that a.c. coupling leads to a restriction of the contrast range in the reproduced picture
The a.c.-coupled video signal settles itself with equal areas on either side of the fixed bias line (see (b)). This results in black level moving to the right on the grey line, which is then reproduced at a higher brightness level than it ought to be.

followed by a large number of grey lines, the resulting black-level moves slowly back from its value under condition A to rest finally as shown for condition B.

Notwithstanding its disadvantage however, a.c. coupling is quite often used, particularly in negative modulation receivers.

Justification for this is argued on two counts:

(a) That the mean level brightness of programmes does not, *on the average*, change very much. Because of the long time constant of the coupling network, therefore, the average charge on the coupling capacitor does not change overmuch so that black-level remains reasonably constant.

(b) That, in any event, even when d.c. coupling *is* used, the form of simple mean-level a.g.c. employed in many receivers creates much the same effect as a.c. coupling anyway.

Despite these arguments, it cannot be denied that a near perfect representation of the transmitted picture can only be obtained if black-level is rigorously maintained constant. The

difference is that which exists between 'optimum' reception and reception which is subjectively considered to be perfectly acceptable and which is economically priced. Such compromises are a common feature in modern life, where complex and costly apparatus is streamlined to be available at realistic domestic prices.

An example of an a.c. coupled vision detector is shown in Fig. 5.12. The time constant of the coupling network is given by:

$$\begin{aligned} C_c \cdot R_g \text{ seconds} &= 0{\cdot}22 \times 10^{-6} \times 820 \times 10^3 \text{ seconds} \\ &= 0{\cdot}22 \times 820 \text{ ms} = 180{\cdot}4 \text{ ms} \end{aligned}$$

The periodic time of one field is 20 ms, and of one complete picture 40 ms. The time constant of the coupling network is thus very long indeed compared with the 64 μs duration of one line, and is $4\frac{1}{2}$ times as long as the periodic time of one picture. It follows that the charge on the coupling capacitor, and hence the positioning of the video waveform beneath the video

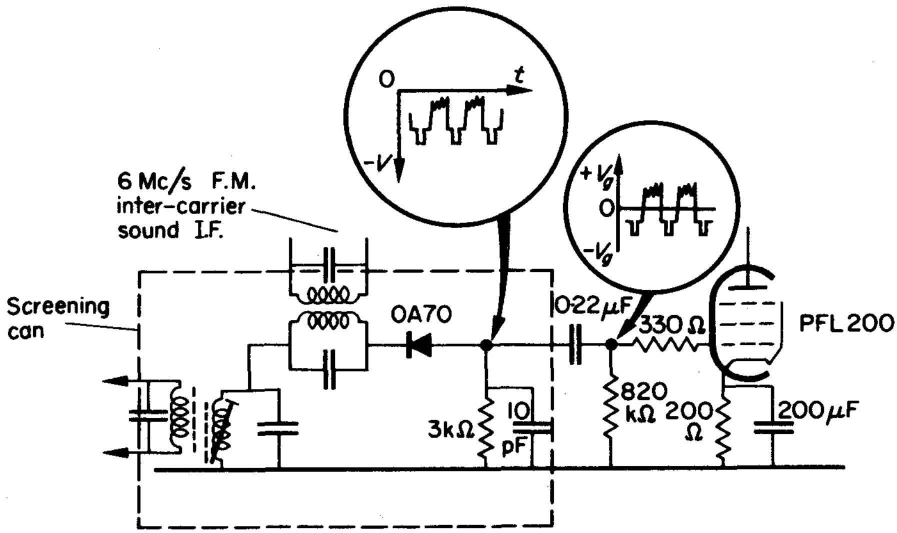



Fig. 5.12 Example of an a.c.-coupled detector in a British 625-line negative-modulation receiver

amplifier's $I_a/V_g$ curve, cannot change significantly from one line to another, so that the relative brightness levels of adjacent lines will be fairly reproduced. Long-term changes in mean level scenic brightness will, however, be followed by the coupling network. The slow resulting drift to and fro of the effective black-level must then inevitably result in a scaling down of the average brightness level in bright scenes but a scaling up of the average brightness level of dark scenes.

**The d.c. restorer diode**

By the use of a d.c. restorer diode it is possible to put back the d.c. component, whilst retaining the advantage conferred by the a.c. coupling, of low anode current under no-signal conditions.

In Fig. 5.13 a restorer diode is shown connected between the grid of the video amplifier and earth. This diode faces its cathode towards the incoming signal. Under these conditions it

conducts on the negative part of the signal. The sync. pulses of the video waveform drive the diode into conduction, so that the coupling capacitor charges up with the plate facing the valve grid becoming positive to chassis. The time constant of the *CR* circuit is of necessity very long compared with the line duration time. After a little while the circuit settles down to a condition such that the diode conducts only during the tips of the sync. pulses, to replace the small discharge which takes place between pulses. Looking outwards from the valve grid, the video signal appears to be applied in series with a steady positive voltage. This has the effect of

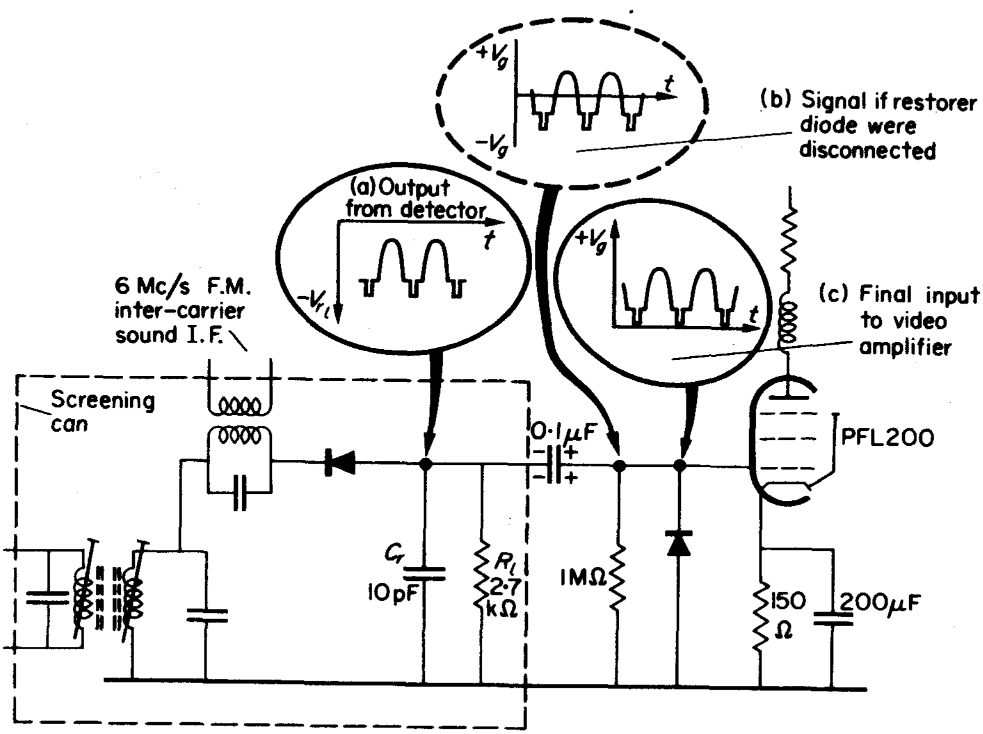



Fig. 5.13 Example of an a.c.-coupled detector followed by a d.c. restorer diode for a British 625-line negative-modulation receiver

lifting the video signal so that, in effect, it is now presented to the valve with the sync. pulses 'sitting upon' the zero voltage chassis potential line. The diode may be regarded as a device which 'does not allow' the video voltage to excurse below the chassis potential. When the tips of the sync. pulses cause the diode to conduct, the capacitor charges and this pushes the net voltage seen by the valve up above chassis potential.

Since all successive lines of video information now rest with the sync. pulses at chassis potential, the effect of the original d.c. component is simulated, and the contrast range is preserved.

The three waveforms shown as insets on Fig. 5.13 serve to revise the principles discussed. Note that:

(a) At the detector anode, the video signal is an entirely negative voltage.
(b) At the grid of the video amplifier, if the restorer diode were not there, the video signal would be set on either side of the zero voltage line in the way described.

(c) With the diode present, the charge on the series coupling capacitor which results from the rectifying action of the diode, 'lifts' the video signal up in potential so that it once more bears a d.c. component. The resulting effective input to the valve appears as a video signal which is now an entirely positive voltage.

It should be noted that in each of the three forms described the *positive-going* nature of the video signal has remained unchanged, and the 'shape' of the waveform is unchanged.

The presence of the d.c. component in a video signal is clearly characterised by all successive sync. pulses being clamped to a fixed voltage level. Whether this particular level is a negative voltage, zero voltage, or a positive voltage, makes no difference to the essential nature of the signal. What it does affect is the required bias conditions of the video amplifier which receives the signal from the detector.

In a negative modulation receiver, the advantage of a.c. coupling the video signal and then restoring the d.c. component is that the required video amplifier bias is converted from the technically undesirable near-zero level necessary, if direct coupling of the all-negative signal from the detector is used, to the safe back-bias needed for working with the all-positive signal which results from the restoration process.

**Dual standards receivers**

Detectors for both positive and negative modulation receivers have been studied. From what has been said it is possible to summarise the factors which have to be taken into account in the detector circuit if a receiver is to be able to switch from one system to another; e.g. to receiver either the British 405-line or the 625-line signals. These are:

(1) The video amplifier must receive video signals of the same polarity on both systems.
(2) Assuming a positive-going video signal input is required by the video amplifier, the signal must be taken from the diode's cathode when handling the positively modulated vision signal, but from its anode when working on the negatively modulated signal.
(3) On the positively modulated signal, the detector video signal output is entirely positive to chassis, so that if d.c. coupling is used the video amplifier must be biased back towards cut-off.
(4) Conversely, with the negatively modulated signal, the video output from the detector is entirely negative to chassis so that with d.c. coupling the video amplifier must have little or no standing bias.
(5) Attention must be paid to the time constant of the detector $C_r R_l$ network. Whilst the vision I.F.s are usually very similar, the video frequency bandwidth may not be. Thus, with the British systems, the I.F.s are often 34·65 Mc/s and 39·5 Mc/s on the 405 and 625-line signals respectively. The corresponding video bandwidths are 3 Mc/s and 5·5 Mc/s respectively. Since the time constant must be short compared with the highest modulating frequency, it is evident that this difference in video bandwidth must not be overlooked.

Assuming d.c. coupling is to be used on both systems, the following possible switching operations may have to be made by the 'systems switch':

(i) *The diode*
(a) A single diode may be reversed by the switch.
(b) Separate diodes may be switched in for each system.

(ii) *The video amplifier bias*

The standing bias must be changed from near cut-off on the positive modulation system to near zero on the negative modulation system.

(iii) *The reservoir capacitor ($C_r$) and load resistor ($R_l$) network*

(a) Separate networks may be switched in for each system.
(b) A single network may be modified for the alternate systems.
(c) A single network, having a compromise time constant, may be employed unchanged on both systems.

Various combinations are clearly possible in the coupling arrangements employed for the two systems, e.g.:

(1) D.C. coupling on both systems.
(2) D.C. coupling with positive modulation but a.c. with negative modulation.
(3) D.C. coupling with positive modulation but a.c. plus partial d.c. coupling on negative modulation. (Partial d.c. coupling is explained under Fig. 5.14 below.)
(4) D.C. coupling on positive but a.c. coupling followed by d.c. restoration on negative modulation.
(5) A.C. coupling on both systems.
(6) A.C. coupling with d.c. restoration on both systems.

Each of these has its own singular advantages and disadvantages. In Britain many different circuit arrangements have appeared in 405/625 receivers. Because of the switching, they tend to look complicated. To understand them it is a great help to draw out the effective individual circuits which result from the alternate switch positions. Two examples of such circuits are illustrated in Fig. 5.14 and Fig. 5.15. When redrawn, the individual circuits appear quite straightforward examples of the circuits discussed in this chapter.

The following points are worth noting in Fig. 5.14:

(a) The diode is reversed by the switch.
(b) The amplifier cathode resistor is reduced from 220 Ω on positive modulation, to 56 Ω on negative modulation.
(c) Although a.c. coupling is employed on the 625-line signal, there is also partial d.c. coupling because the 0·1 μF capacitor is bridged by a 1 MΩ resistor. If this resistor is considered together with the 220 kΩ grid resistor, it will be noted that about one-sixth of the negative d.c. voltage output from the detector, is passed on to the valve. This gives a measure of automatic adjustment of the working point beneath the amplifiers $I_a/V_g$ curve. On very bright and strong signals the working point will tend to move backwards from the nominal position determined by the 56-Ω cathode resistor.
(d) Two small H.F. compensation coils $L_7$ and $L_8$ are fitted in series with the 2·7 kΩ load resistor when switched to 625. This enables the circuit to handle the wider frequency range of the 625-line video signal.
(e) As is the case with many vision detectors, the diode is enclosed in the last I.F. screening can to avoid radiation of interference back to the tuner unit.
(f) In this particular circuit a part only of the secondary winding of the I.F. transformer, is coupled to the primary.
(g) In the 405-line position, the video amplifier cathode circuit includes a 3·5 Mc/s inter-I.F. beat frequency rejector circuit. This unwanted component is developed across the

Effective circuit on the 405-line positively-modulated system

Effective circuit on the 625-line negatively-modulated system

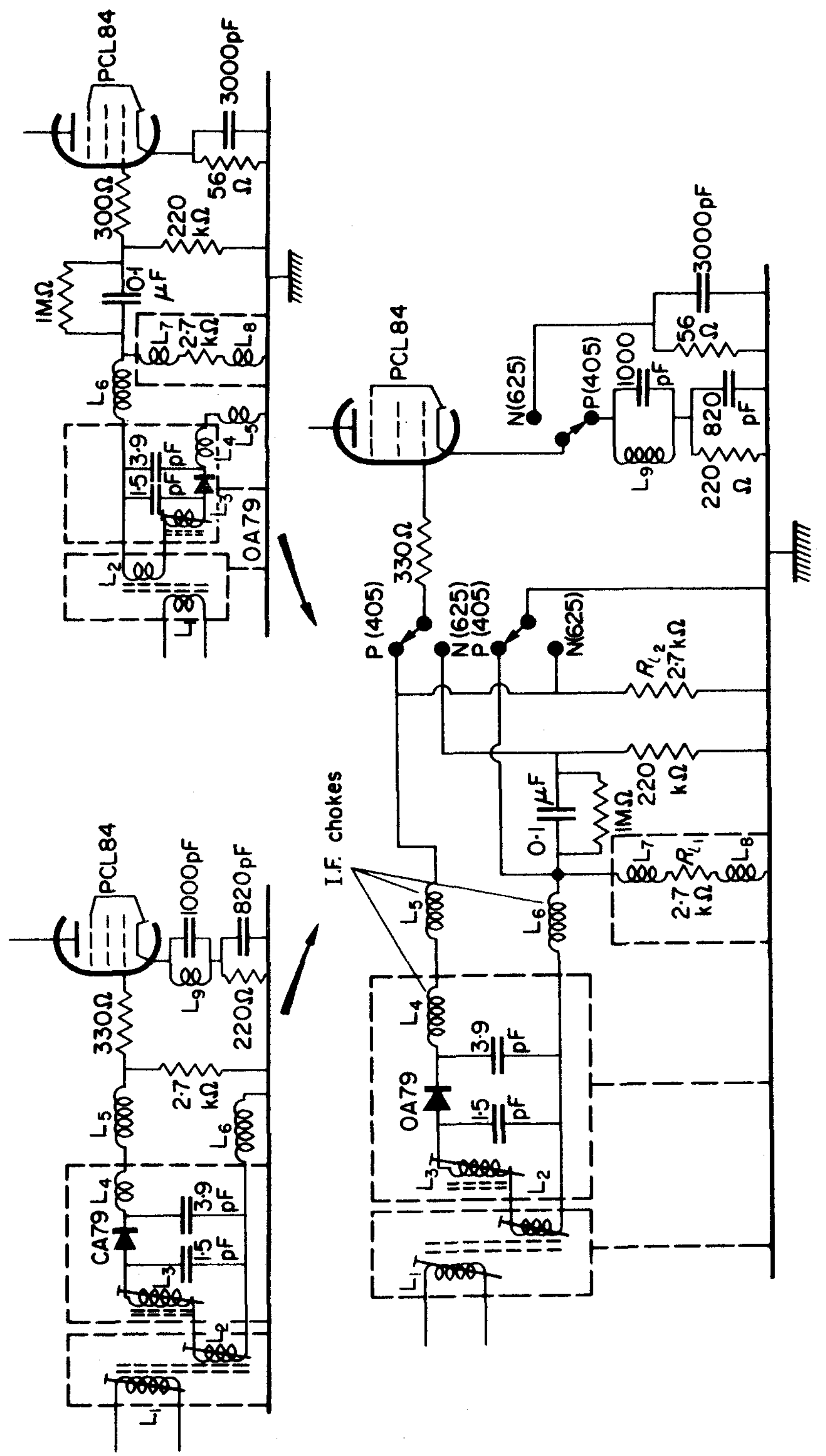



Fig. 5.14 Detector circuit in a dual-standards receiver designed to work on the British 405-line positively-modulated and 625-line negatively-modulated signals

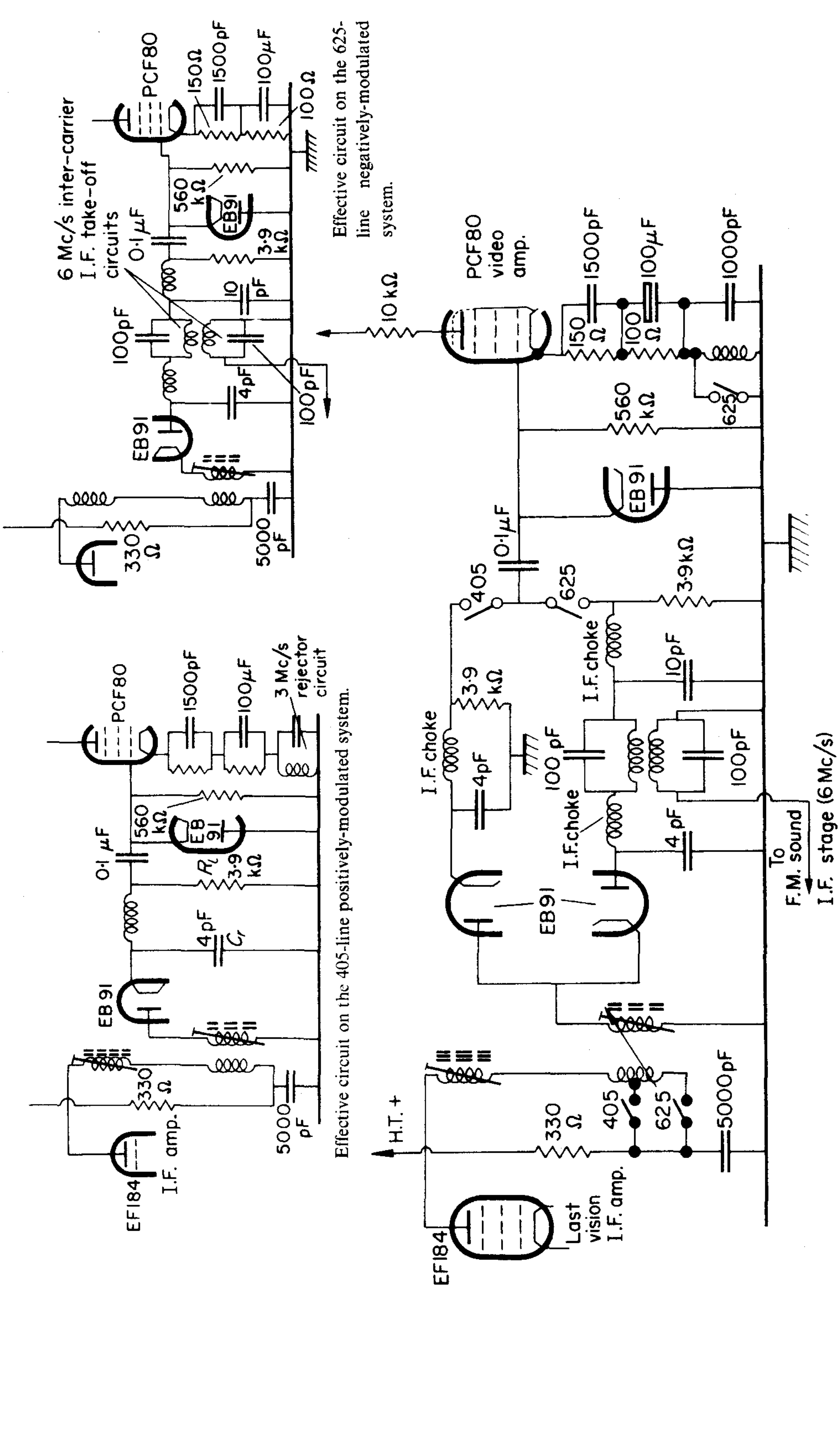



Fig. 5.15 Dual-standards vision detector using a.c. coupling followed by d.c. restoration on both positively- and negatively-modulated systems

tuned-circuit where it gives rise to a high level of negative feedback at 3·5 Mc/s. The result is that this beat frequency does not go forward to the c.r.t. where it would cause a fine grain-like interference pattern on the raster.

The circuit of Fig. 5.15 uses a.c. coupling with d.c. restoration on both systems. As demonstrated in the diagrams of Fig. 5.10(c) this results in identical inputs to the video amplifier on both systems. The switching is therefore more simple and the main operation is to connect the coupling capacitor to either the positive or negative modulation detectors on 405 or 625 respectively. On 625-lines the detector filter circuit is more complex since it includes the F.M.-sound 6 Mc/s inter-carrier I.F. take-off circuit. A further switching operation inserts a 3·5 Mc/s inter-I.F. beat frequency rejector circuit into the video amplifier cathode on the 405-line signal. Finally, the coupling factor of the I.F. transformer is modified to give optimum matching and bandwidth on the two systems.

**Transistorised circuits**

With transistorised sound receivers for normal broadcast reception, modifications are needed to adjust the output impedance of the detector to match into a transistor instead of a valve. For example, in the case discussed earlier in this chapter, values for $C_r$ and $R_l$ of 100 pF and 500 kΩ were used. With transistor receivers, the value of $R_l$ must be decreased by a factor of 100, to 5 kΩ, and to retain the same time constant of 50 μs, the capacitor is increased by the same factor of 100, to 0·01 μF.

In transistorised television receivers, however, the vision detector remains very much the same as it is in valve television receivers. This is because the low value of load resistor employed in vision detectors is already of the right order for matching into a transistor video amplifier (i.e. a few thousand ohms).

Due attention must be paid in the arrangement of the d.c. coupling network, to the required forward bias for the transistor base-emitter junction.

To minimise the shunting effect on the detector circuit, vision detectors often feed the video amplifier via an emitter-follower stage. Fig. 5.16 shows a typical circuit. Like a valve cathode follower, the emitter-follower offers a comparatively high input impedance and low output impedance. Typical figures for emitter-follower video driver stages show input resistances of the order 25 kΩ to 50 kΩ and output resistances of 25 Ω to 100 Ω.

In the circuit shown, the transistor is a p.n.p. type. It has to conduct throughout the video signal waveform and must be so biased that the signal fits entirely beneath the input ($I_b/V_b$) characteristic. For this to be so, the transistor must have a forward bias on the base emitter junction; i.e. the base must be held at a negative potential relative to the emitter. This is arranged by the normal potential divider method. The detector 5·6 kΩ load resistor is made to form part of this divider network. The actual detector circuit is seen to be of the same form as described for valve receivers. The diode is connected to deliver a positive going signal to the transistor base-emitter junction. The positive going video signal now causes a reduction in collector current. This is because the net base-emitter voltage is the algebraic sum of the fixed negative bias and the instantaneous value of the positive going signal. From this it follows that the forward bias must set the operating point well forward under the $I_b/V_b$ curve and hence well forward under the $I_c/I_b$ transfer characteristic. The applied signal then excurses 'backwards' under the curve.

It should be noted that peak-white corresponds to the minimum net negative voltage on the base and hence the minimum value of collector current. As $I_c$ reduces, however, so also does the

volts drop across the emitter load resistor, and this results in the emitter voltage moving nearer to the battery positive potential, i.e. becoming *less negative* to chassis. To sum up, in the diagrams shown, a positive going input signal causes a *reduction* in collector current and this in turn gives rise to a positive going output at the emitter.

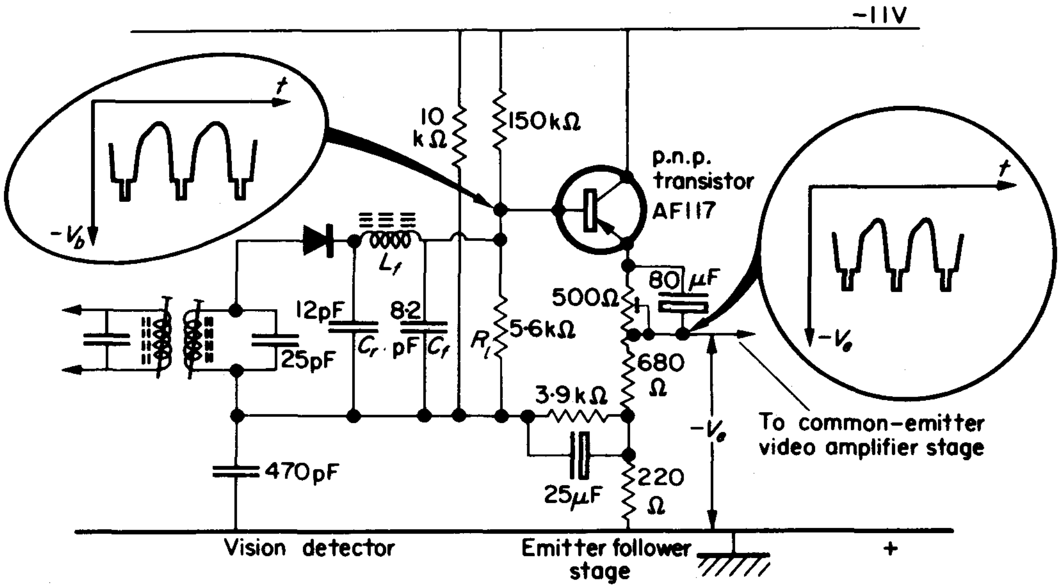



Fig. 5.16 Vision detector in a positive-modulation transistorised television receiver

The question of signal polarities in transistor circuits and of the graphical representation of operating conditions is dealt with in some detail in Chapters 8 and 10 under the sections on transistorised video amplifiers and transistorised sync. separators respectively. As with valve circuits, however, it is only necessary to work backwards from the c.r.t. to determine what polarity of signal is required from the vision detector. In the example shown in Fig. 5.16 the emitter follower feeds the base of a single common-emitter video amplifier stage, and the c.r.t. cathode receives a negative-going video signal from the collector of this video amplifier.

It should be noted in this diagram that the emitter load consists of the series combination of the 680 Ω and 220 Ω resistors, together with a part of the pre-set 500 Ω variable resistor. The latter allows the total emitter load resistance to be adjusted for optimum working conditions.

The 3·9 kΩ resistor forms part of the necessary potential divider chain to fix the forward bias at the correct level. If it were combined with the 5·6 kΩ resistor, the detector load resistance would then be too high. The 3·9 kΩ is therefore decoupled so that it carries out its d.c. function, without interfering with the a.c. performance.

# 6

# Video Amplifiers: General Principles

## Performance requirements

(a) *Gain.* Typically the gain of the video amplifier stage in valve receivers is of the order of 10 to 25 times, e.g. under normal working conditions the video signal input to the video amplifier from the detector may be of the order of 2 volts amplitude, and the output signal delivered by the video amplifier to the c.r.t. of, say, 50 volts amplitude, showing a gain of 25 times.

Table 6.1 shows examples of the video signal amplitudes found in some typical receivers when producing a well-contrasted picture.

**Table 6.1**

*Showing examples of typical video amplifier working conditions, measured in each case with the receiver producing a well-contrasted picture*

| Video signal input to video amplifier from detector Volts: peak-to-peak | Video signal output from video amplifier Volts: peak-to-peak | Implicit gain of video amplifier | Video signal input to c.r.t. | Video amplifier load resistor |
|---|---|---|---|---|
| 2·5 V | 50 V | 20 | 47 V | 4·7 kΩ |
| 6·0 V | 50 V | 8·3 | 48 V | 9·1 kΩ |
| 3·0 V | 42 V | 14·0 | 40 V | 6·8 kΩ |
| 5·0 V | 72 V | 14·4 | 72 V | 4·4 kΩ |
| 6·5 V | 82 V | 12·6 | 75 V* | 4·7 kΩ |
| 3·0 V | 65 V | 22 | 62 V | 5·1 kΩ |

* High level contrast control used in this receiver.

(b) *Bandwidth.* Ideally the video amplifier's response curve should be linear from d.c. (i.e. 0 c/s) up to the highest video frequency employed by the system.

Table 6.2 below shows the desirable video amplifier bandwidths for some of the principal television systems.

**Table 6.2**
*Examples of desirable video amplifier bandwidths*

| System | Video amplifier bandwidth |
|---|---|
| British 625-line | d.c. to 5·5 Mc/s |
| British 405-line | d.c. to 3 Mc/s |
| C.C.I.R. 625-line | d.c. to 5 Mc/s |
| American 525-line | d.c. to 4·2 Mc/s |
| French 819-line | d.c. to 10·4 Mc/s |

(c) *Phase response.* The relative phases of all the frequency components present in the video signal must be preserved.

If there is *no* phase shift at *any* frequency, then of course this requirement is satisfied.

In amplifiers of several stages, with reactive components (e.g. coupling capacitors) present in the coupling networks, phase shift is inevitable. This is obvious since the phase angle, together with the circuit impedance, varies with frequency.

Phase shift is not important in A.F. amplifiers because the ear does not detect changes in the relative phases of the various frequency components present in a given sound. It *is* important in video amplifiers since phase shift implies 'time' shift, which in turn implies positional shift in the reproduced visual image. Fortunately, the form of single stage or two stage, video amplifiers used in most television receivers, provided they have an adequate frequency bandwidth, do not present significant phase shift difficulties. However, there are occasions when a knowledge of the problems of phase shift is important, and it is instructive to consider the question a little further.

If phase shift *is* present, then the requirement for zero distortion at the output of the amplifier is that the amount of phase shift introduced by the circuit must be *proportional* to *frequency.*

Thus for zero phase distortion there must be either:

(1) No phase shift at all at any frequency; or
(2) an amount of phase shift which is *proportional* to frequency.

It is easy to fall into the error of thinking that if there is phase shift then it should be the same for all frequencies. The diagrams in Fig. 6.1 demonstrate that this is not so, and that the statement (2) is correct.

The input to the amplifier is assumed to consist of three separate frequency components. For convenience they are shown as a fundamental A of frequency $f$ c/s, a second harmonic B of frequency $2f$ c/s, and a third harmonic C of frequency $3f$ c/s.

For faithful reproduction, these frequencies must come out of the amplifier with:

(a) The correct relative amplitudes.
(b) The same relative phases.

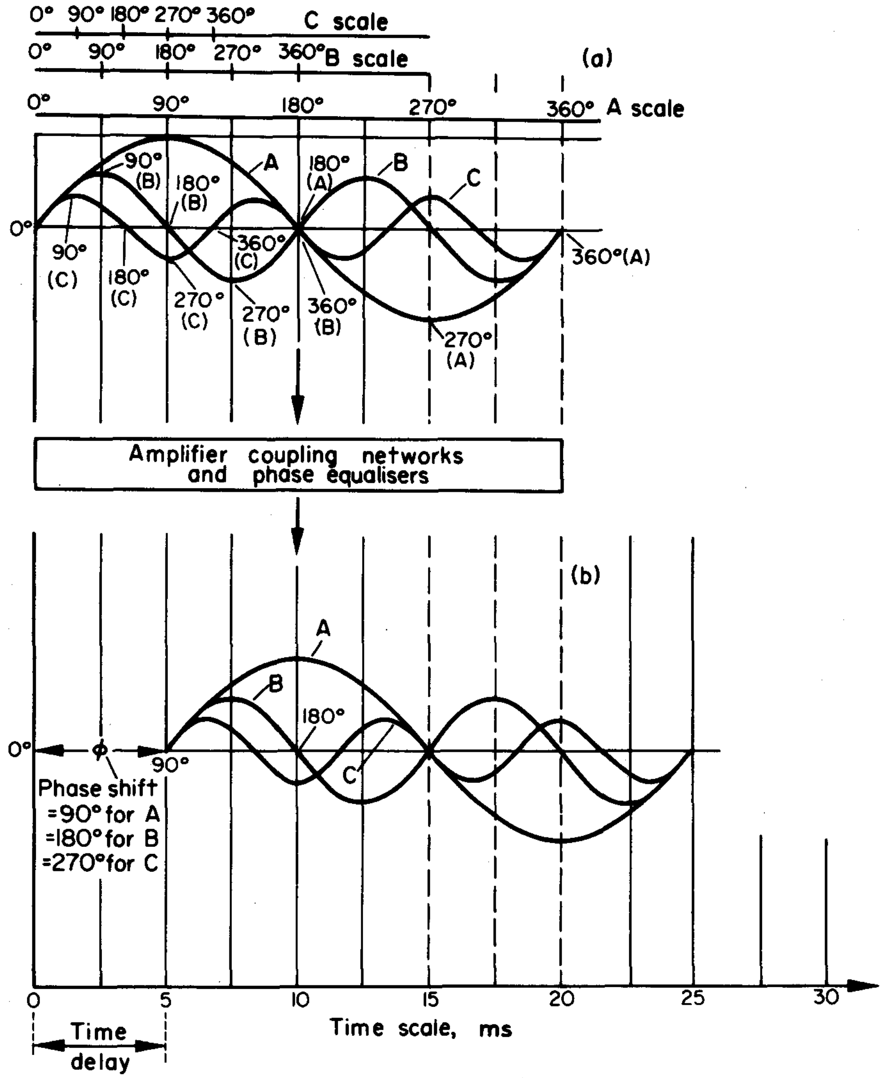



Fig. 6.1 Showing that, for zero phase distortion, the phase shift through an amplifier must be proportional to frequency (*or* zero at *all* frequencies)

(a) This diagram shows the assumed phase relationships of three separate *input* components to an amplifier:

$$A = f_0 \text{ c/s} (= 50 \text{ c/s})$$
$$B = 2f_0 \text{ c/s} (= 100 \text{ c/s})$$
$$C = 3f_0 \text{ c/s} (= 150 \text{ c/s})$$

(b) In this output signal, the *relative phases* are still the same. For this to be so, if $A$ suffers 90° phase shift, then $B$ and $C$ must experience 180° and 270° phase shift respectively.

If the $x$-axis is scaled in time rather than angular displacement, then the same scale serves all the components. To retain the correct 'phase' relationships, each must suffer the *same time shift* (in this case 'delay'). But, to $A$ at 50 c/s, 5 ms represents 1/4 of the periodic time of 20 ms, and this in terms of angular displacement is $360/4 = 90°$. Similarly, to $B$ at 100 c/s, 5 ms is 1/2 of the periodic time of 10 ms and this represents a phase shift of $360/2 = 180°$. To $C$, at 150 c/s, 5 ms $= 3/4$ of the periodic time of 6·67 ms which represents a phase shift of 270°.

Suppose the coupling networks shift the phase of A by a total of 90°. A study of the diagram shows that if B and C are to retain their position (phases) relative to A, then it is necessary that:

(1) B is shifted by 180°;
(2) C is shifted by 270°.

Thus if the fundamental frequency of $f$ c/s is shifted by 90°, then a frequency of $2 \times f$ must be shifted by $2 \times 90 = 180°$ and the third harmonic of frequency $3 \times f$ must be shifted by $3 \times 90° = 270°$; i.e. the phase shift must be *proportional to frequency*.

The diagram shows that to keep the components in their correct relationship one with another, they must all be shifted the same *distance* along the graphical $x$-axis *but* this does not imply equal degrees of phase shift, since the angular scale along the $x$-axis is of course different for each frequency.

Often *time*, rather than *phase*, is considered in order to arrive at the same conclusion.

Suppose a certain frequency component '$f$' is shifted back in phase by $\phi°$. If the time duration of one cycle at the frequency $f$ c/s is $t$ seconds, then, since one cycle of 360° takes $t$ seconds, the time delay corresponding to a phase shift of $\phi°$ must be $\left(\frac{\phi}{360} \times t\right)$ seconds.

But $t = \frac{1}{f}$ and hence the time delay is $\left(\frac{\phi}{360} \times \frac{1}{f}\right)$ seconds.

If a frequency of $2f$ is now considered, and the phase shift is held constant at $\phi°$, the time delay will now be $\left(\frac{\phi}{360} \times \frac{1}{2f}\right)$ seconds, or one *half* what it was. For freedom from distortion, all components must suffer the *same* 'time shift'. To achieve this, if $f$ increases to $2f$, the phase shift $\phi°$ must increase to $(2 \times \phi)°$, so that the time delay is then given by:

$$t = \left(\frac{2\phi}{360} \times \frac{1}{2f}\right) \text{ seconds} = \left(\frac{\phi}{360} \times \frac{1}{f}\right) \text{ seconds}$$

which is the same as before.

The conclusion, once again, is that phase shift must be proportional to frequency if all frequency components are to 'arrive' at the output of the amplifier with the same relative phases that they possess at the input.

In Fig. 6.1(b) the $x$-axis has been scaled in time, rather than phase angle. The same scale now obviously serves all the various frequencies.

To retain their correct relative positioning (phases) all must undergo the same time delay. Waveform A, assumed to be a 50 c/s waveform, is shown to be delayed by 5 ms, which represents a phase shift of 90°. But, if waveform B, which is at 100 c/s, is also delayed by 5 ms, this represents the time of one complete $\frac{1}{2}$-cycle and this in turn involves a phase shift of $2 \times 90 = 180°$. Similarly, 5 ms to waveform C represents $\frac{3}{4}$-cycle; i.e. a 270° phase shift.

In multi-stage amplifiers for video frequency working, it is necessary to employ 'phase equalising' circuits to give the required proportionality between phase shift and frequency. The usual form of phase shift distortion, inherent in R.C. coupled amplifiers, is such that low frequencies develop a leading phase angle relative to middle range frequencies, and high frequencies develop a corresponding phase lag.

Before leaving this subject a common source of misunderstanding should be noted. It is often said that a normal valve amplifier (or transistor C.E. amplifier) introduces 180° phase shift. This is not true. What happens is that the signal is 'inverted'. With a sine wave, inversion

does in fact produce the same effect as 180° phase shift, *but* this is only true because the waveform is symmetrical about the $x$-axis. In the case of a complex waveform such as a video signal, inversion is obviously *not* the same as phase shift. The point is illustrated in Fig. 6.2.

The simultaneous inversion of all signal frequency components does not therefore effect their relative phases. In resistive loaded amplifiers it is the presence of reactive components in the coupling circuitry (not forgetting stray shunt capacitance) which affects phase response.

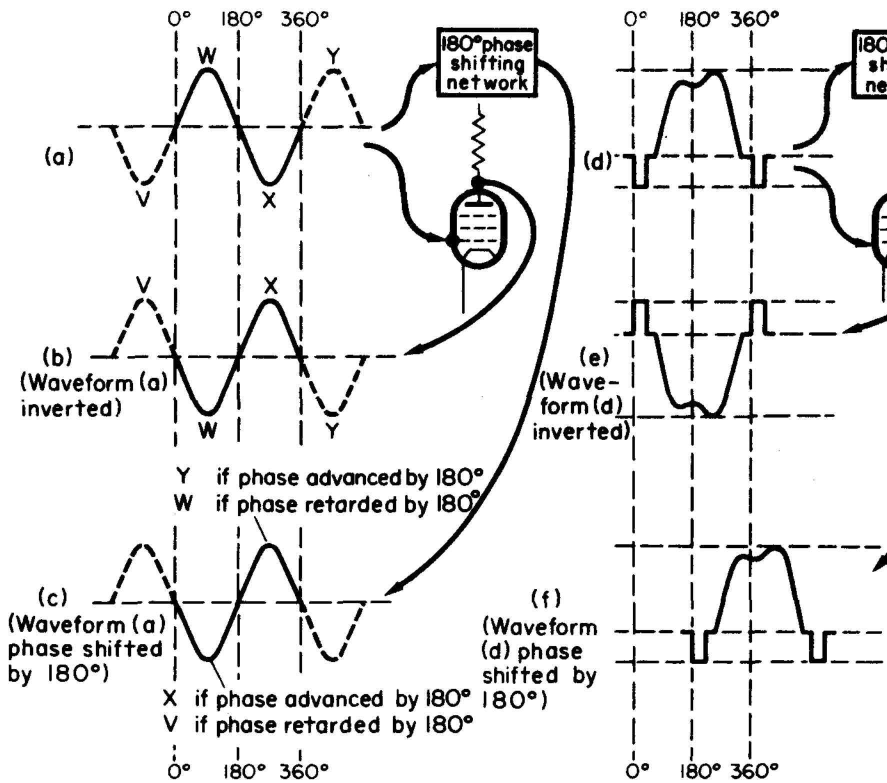



Fig. 6.2 Diagrams showing the distinction between signal inversion and 180° phase shift

*Note* that:

(i) A valve only produces the apparent effect of 180° phase-shift when the signal is entirely symmetrical about the $x$-axis (e.g. a sine-wave).

(ii) A phase shifting network moves the signal horizontally along the $x$-axis. The effect can be seen by a study of the sine-wave peaks labelled $v$, $w$, $x$, $y$ in Figs (a) and (c), and by the video signal at (d) and (f).

(iii) A valve does not move the signal horizontally; it merely inverts it. (Again, see the labels $v$, $w$, $x$, $y$ of (a) and (b).)

(iv) A transistor amplifier in the common emitter mode produces the same effect as the valve amplifier shown.

## Wideband amplifier theory

A video amplifier is an example of a wideband amplifier. Such amplifiers are best studied by dividing the frequency range handled, into three regions; low frequencies (L.F.), medium frequencies (M.F.) and high frequencies (H.F.). These terms are relative to the particular range covered, and what is considered to be a high frequency in one application may of course be only a medium frequency or even a low frequency in another case.

In general, a single-amplifier stage is a four-terminal circuit. Usually one input and one output terminal are common. The coupling circuitry to the next stage must be included in the amplifier circuit, and the input terminals to the next stage thus form the output terminals of the stage under consideration. In this way the influence upon performance of every part and component of the complete circuit is taken into account.

Fig. 6.3(a) shows the basic circuit to be considered. $C_c$ is for the moment assumed to be necessarily included for d.c. blocking purposes. Its value will be chosen to offer low reactance

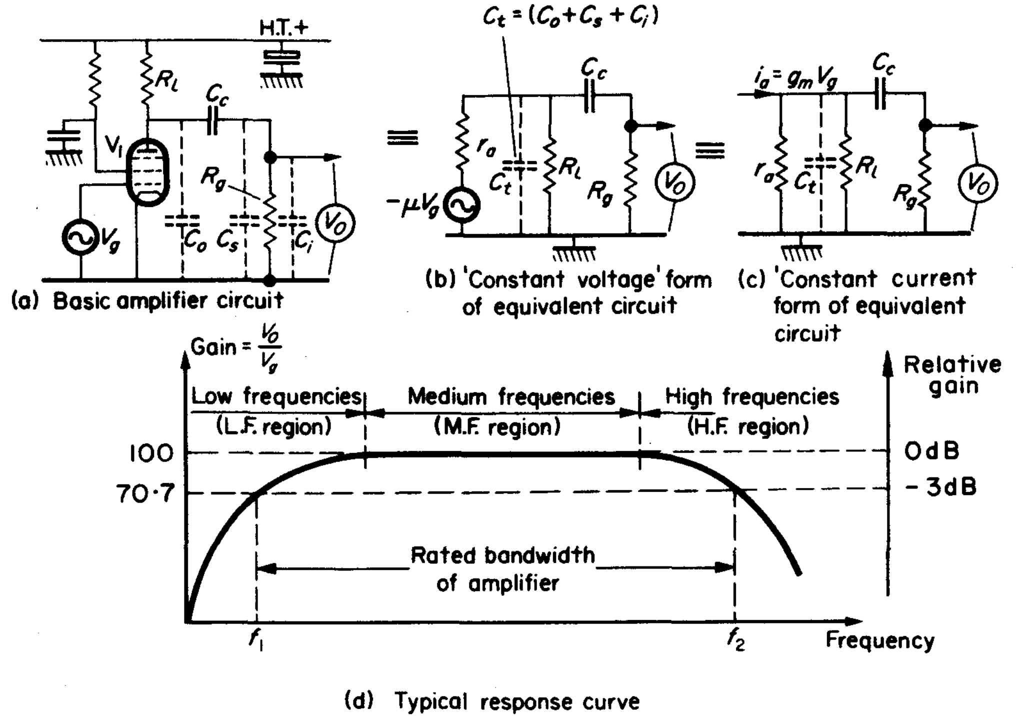



Fig. 6.3 Showing the basic form of a voltage amplifier with equivalent a.c. circuits and the general shape of the gain/frequency response curve

over the pass-band of the amplifier. $C_o$ represents the output capacitance of $V_1$, $C_s$ the stray shunt circuit capacitance of the coupling components and wiring, whilst $C_i$ is the input capacitance to the next stage.

Fig. 6.3(b) shows the equivalent a.c. circuit of the stage.* Looking towards $V_1$ from $C_s$, $C_o$ appears in series with $C_c$ and since $C_c$ is much larger than $C_o$, the effective shunt capacitance 'seen' by $C_s$, is merely $C_o$. Thus $C_o$ is effectively in parallel with $C_s$ and $C_i$. In the diagram $C_t = C_o + C_s + C_i$. Fig. 6.3(c) shows the alternative 'constant current' form of the equivalent circuit. This is sometimes more convenient to use.

Suppose a constant voltage variable frequency oscillator is now connected to the input terminals and the output voltage developed across $R_g$ is measured at various frequencies throughout the range of the amplifier. Typically, the frequency response curve obtained is of

* For those not familiar with equivalent circuits, a simple explanation is given in the Appendix, p. 222.

the form shown in Fig. 6.3(d). This shows a central flat M.F. region where the gain is constant, bordered by regions at L.F. and H.F. where the gain progressively falls off. The gain in the flat M.F. central region is used as the reference level against which the gain at other frequencies is compared.

The bandwidth of an amplifier is measured between the two frequencies $f_1$ and $f_2$ at which the gain falls to 70·7% (i.e. 0·707 or $1/\sqrt{2}$) of its central (M.F.) value.

In terms of decibels a voltage-gain ratio of unity equals 0 dB, whilst a ratio of $1/\sqrt{2}$ equals −3 dB. The flat M.F. level of gain is therefore often shown as 0 dB, with the relative gain at $f_1$ and $f_2$ showing as −3 dB. Sometimes, since a drop of 3 dB corresponds to a halving of the 'power' output, these frequencies are referred to as the 'half-power' points of the response curve.

In studying amplifier performance, and in deducing what must be done to extend the bandwidth, it is convenient to focus attention on these two frequencies. Initially, an expression for the gain in the M.F. region is obtained. Thereafter, the gain at H.F. and the gain at L.F. may each be expressed *in terms* of the gain at M.F. It is then a simple matter to deduce the frequencies at which the gain falls from 'gain at M.F.' to 'gain at M.F.'$/\sqrt{2}$, and thence to see what has to be done to extend the bandwidth.

Notice that to extend the bandwidth involves making either $f_2$ higher or $f_1$ lower, or both.

*Gain at M.F.* From inspection of the circuit it is obvious that $C_t$ is the chief reason for the fall-off of gain at H.F. and $C_c$ for the fall-off at L.F.

Clearly, in the central M.F. region, where the gain is constant, both $C_t$ and $C_c$ have a negligible effect. If this were not so, the response curve would not *be* flat in this region.

The equivalent circuit for medium frequencies may therefore be redrawn without $C_c$ and $C_t$, as shown in Fig. 6.4. Writing $R_t$ for the combined resistance of $r_a$, $R_l$ and $R_g$ in parallel now

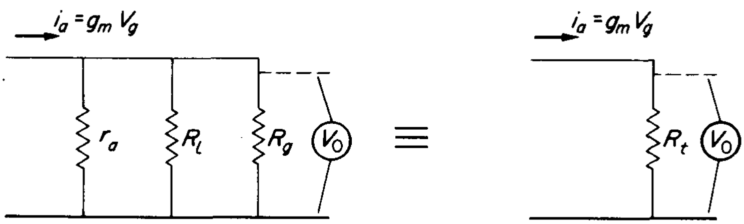



Fig. 6.4 Showing the equivalent circuit at medium frequencies (M.F.) where both the coupling capacitor $C_c$ and the total shunt capacitance $C_t$ have negligible effect

gives a very simple equivalent circuit, from which an expression for the gain at M.F. may be found.

Thus
$$V_o = i_a R_t$$
where $i_a$ is the a.c. component of anode current.

But
$$i_a = g_m V_g$$
$$\therefore\ V_o = g_m . V_g . R_t$$

Hence:
$$\text{Gain at M.F.} = \frac{V_o}{V_g} = g_m R_t \qquad (1)$$

*Gain at H.F.* Here the series coupling capacitor $C_c$ is again of no consequence, but the shunt capacitance $C_t$ now affects the net load impedance and must appear in the equivalent circuit

as shown in Fig. 6.5. In the M.F. range the reactance of $C_t$ is high compared with the resistance $R_t$ and its presence does not modify the effective load impedance. At H.F., however, the falling reactance of $C_t$ lowers the net impedance and causes a drop in gain.

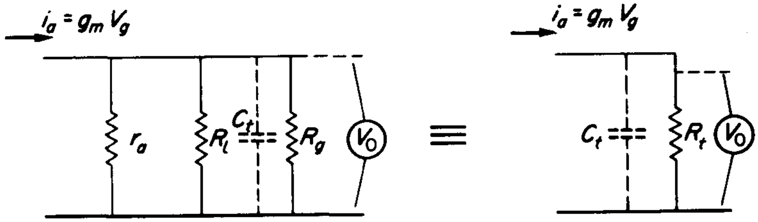



Fig. 6.5 Showing the equivalent circuit at high frequencies (H.F.), where $C_c$ has negligible effect but $C_t$ becomes increasingly significant

The same approach as before gives an expression for the gain.

Thus $$V_o = i_a Z_t = g_m . V_g . Z_t$$

(where $Z_t$ is the parallel impedance of $C_t$ and $R_t$)

$$\therefore \text{ Gain at H.F.} = \frac{V_o}{V_g} = g_m Z_t$$

Manipulation of this expression yields:

$$\text{Gain at H.F.} = \frac{g_m R_t}{\sqrt{1+(\omega C_t R_t)^2}}$$

i.e. $$\text{Gain at H.F.} = \frac{\text{Gain at M.F.}}{\sqrt{1+(\omega C_t R_t)^2}} \quad . \quad . \quad . \quad . \quad . \quad (2)^*$$

Obviously, the gain at H.F. falls to $1/\sqrt{2}$ of the gain at M.F. when $\omega C_t R_t = 1$. But the frequency at which this happens has been defined as $f_2$, the upper half-power frequency. Hence $f_2$ may be expressed in terms of circuit constants.

Thus $$\omega_2 C_t R_t = 1$$
$$\therefore \; 2\pi f_2 C_t R_t = 1$$

and $$f_2 = \frac{1}{2\pi C_t R_t} \quad . \quad . \quad . \quad . \quad . \quad . \quad (3)$$

Before going on to study what can be done to increase $f_2$ (thereby extending the bandwidth), the corresponding expression for the 'Gain at L.F.' and for the lower half-power frequency $f_1$ will be obtained. The problem of extending the bandwidth in both directions may then be studied with all the relevant data to hand.

*Gain at L.F.* In this case the coupling capacitor $C_c$ has a dominant influence on the circuit, whilst $C_t$ is of course of no consequence. The effective equivalent circuit is shown in Fig. 6.6. As the frequency falls, the reactance of $C_c$ rises and an increasing proportion of the signal is 'lost' across this capacitor.

* The working is shown in the Appendix at the end of the book for those who wish to follow it.

Analysis of this circuit yields*:

$$\text{Gain at L.F.} = \frac{\text{Gain at M.F.}}{\sqrt{1+\left(\dfrac{1}{\omega C_c R'}\right)^2}}$$

where
$$R' = \frac{r_a R_l}{r_a + R_l} + R_g$$

The gain at L.F. falls to $1/\sqrt{2}$ of the gain at M.F. when $1/(\omega C_c R') = 1$. But the frequency at which this happens has been specified as $f_1$. Thus:

$$\frac{1}{2\pi f_1 C_c R'} = 1$$

$$\therefore f_1 = \frac{1}{2\pi C_c R'}$$

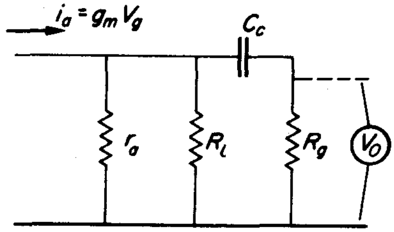



Fig. 6.6 Showing the equivalent circuit at low frequencies (L.F.), where $C_t$ has negligible effect, but $C_c$ becomes increasingly significant

*Extending the bandwidth*

To extend the bandwidth, $f_2$ must be made higher and $f_1$ lower. Dealing first with the H.F. end, a study of the expression for $f_2$ shows that $f_2$ may be made higher by:

(a) making $C_t$ as low as possible;
(b) decreasing $R_t$.

Having taken all possible steps to keep $C_t$ as low as possible, it remains to reduce $R_t$ by decreasing $R_l$. Obviously, decreasing $R_t$ decreases the gain at M.F., so that the price which is paid for shifting $f_2$ upwards is a general overall lowering of gain. In video amplifiers, where $f_2$

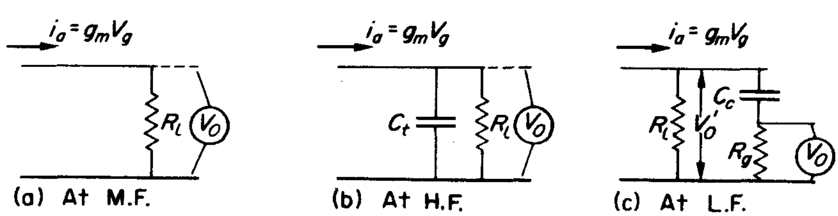



Fig. 6.7 Equivalent circuits at M.F., HF. and L.F., when $R_l$ is very low in value compared with $r_a$ and $R_g$

* See Appendix at the back of the book.

is required to be of several Mc/s, $R_l$ has to be reduced to a point where it is so much less than $r_a$ (and the input resistance of the c.r.t. $R_g$) that $R_t$ becomes sensibly equal to $R_l$. The effective equivalent circuits at M.F., H.F. and L.F. then take the simple form shown in Fig. 6.7. In this case, $R_l$ may be substituted for $R_t$ in the expressions studied. These then become:

(a) Gain at M.F. $= g_m R_l$

(b) Gain at H.F. $= \dfrac{g_m R_l}{\sqrt{1+(\omega C_t R_l)^2}}$

(c) $f_2 = \dfrac{1}{2\pi C_t R_l}$

(d) Gain at L.F. $= \dfrac{\text{Gain at M.F.}}{\sqrt{1+\dfrac{1}{(\omega C_c R_g)^2}}}$ $\quad\left(\text{Since } R' = \dfrac{R_l r_a}{R_l+r_a}+R_g \quad \therefore\ R' \simeq R_g\right)$

(e) $f_1 = \dfrac{1}{2\pi C_c R_g}$

*Example.* A video amplifier uses a pentode of $g_m = 6$ mA/V. The anode load resistor is of 5 kΩ, and the total shunt capacitance across the anode circuit is 20 pF. Find:

(a) The gain of the stage at M.F.
(b) The frequency $f_2$ at which the gain falls by 3 dB.
(c) The answers to (a) and (b) if $R_l = 2{\cdot}5$ kΩ.

*Answers*

(a) Gain at M.F. $= g_m R_t = \dfrac{6}{1000} \times 5000 = 30$

(b) $f_2 = \dfrac{1}{2\pi C_t R_t} = \dfrac{1}{2\pi \times 20 \times 10^{-12} \times 5000}$

$$= \frac{10^{12}}{2\pi \times 10^5} = \frac{10^7}{2\pi} \simeq 1{\cdot}6 \text{ Mc/s}$$

(c) If $R_l = 2{\cdot}5$ kΩ, then:

$$\text{Gain at M.F.} = \frac{6}{1000} \times 2500 = 15$$

$$f_2 = \frac{10^{12}}{2\pi \times 20 \times 2500} \simeq 3{\cdot}2 \text{ Mc/s}$$

The answers are instructive. The figure for $C_t$ is realistic and it is seen that even with $R_l$ down to 2·5 kΩ, the bandwidth only extends up to 3·2 Mc/s. It is possible, however, to extend the bandwidth by various means which avoid undue reduction in the value of $R_l$. Some of these methods are studied below under 'H.F. Compensation'.

A further matter must be mentioned before these techniques are studied. This concerns the anode current of the video amplifier valve. A certain peak-to-peak video signal voltage output is required from the video amplifier in order to drive the c.r.t. fully and set up a well-contrasted

picture. Typically the signal has to be of the order of 60 volts peak-to-peak. Note that by peak-to-peak is meant the voltage change between peak-white level and the tips of sync. pulses. Of this some 40 volts is picture content measured between the black and peak-white levels.

As $R_l$ is decreased, the peak-to-peak anode current variation needed to establish this drive voltage increases. For example, if $R_l$ is 4 kΩ, the required anode current change to set up a 60 volt peak-to-peak video signal is $(60 \times 10^3)/(4 \times 10^3) = 15$ *m*A.

With $R_l = 2{\cdot}5$ kΩ, the anode current change becomes 24 mA.

These figures explain why it is that power valves are often found in the video amplifier stage. However, this is not always done; sometimes R.F. pentodes are used. Since R.F. pentodes have a lower output capacitance $C_o$, a higher value of $R_l$ may be used for the same bandwidth and this in turn limits the maximum necessary anode current.

## H.F. compensation techniques

These circuit techniques are designed to compensate for the fall-off in H.F. response which takes place due to the presence of the shunt capacitance $C_t$.

In general the addition of an H.F. compensation circuit to an amplifier may be made to give rise to one of the following changes in its performance:

(1) An *increased* bandwidth (i.e. higher $f_2$) at the *same* medium and low frequency gain as before.
(2) The *same* bandwidth with *higher* overall gain.
(3) A change intermediate between (1) and (2) such that both the bandwidth and the overall gain increase, though to a less extent than the respective individual increases implied under (1) and (2).
(4) An increased bandwidth with *less* gain at medium and low frequencies.

Obviously, whether or not increased overall gain is sought, or whether or not a sacrifice in overall gain is tolerated, depends upon the inherent gain of the amplifier before the compensation circuit is added. The primary purpose is to push up the $-3$ dB frequency $f_2$, and examples of each of the four possibilities mentioned above are found in practice.

### *Method 1. The cathode follower*

By inserting a cathode follower stage between the video amplifier and the c.r.t., the input capacitance ($C_i$) to the tube, and a substantial part of the shunt capacitance ($C_s$) of the wiring, is removed from the video amplifier anode circuit. The decrease in $C_t$ makes possible a higher value of $R_l$ and an R.F. pentode may then more readily be used as the video amplifier.

The basic circuit arrangement is illustrated in Fig. 6.8.

A cathode follower has the following properties:

(a) A gain of less than unity. (N.B. The nature of the circuit results in 100% negative feedback being applied.)
(b) A very high input impedance.
(c) A very low input capacitance.
(d) A very low output impedance. (This is approximately equal to $1/g_m$ Ω.) For example, a valve of $g_m = 5$ mA/V, used as a C.F., presents an output impedance of approximately

$$\frac{1}{g_m} = \frac{1}{5 \times 10^{-3}} = 200\ \Omega$$

The influence of these properties upon the performance of a video amplifier may be examined one at a time.

(a) *Gain.* The fact the C.F. stage has a gain of slightly less than unity is of no consequence. Its presence allows for a higher value of $R_l$ in the video amplifier and so it does in fact confer higher overall gain. It should be noted that the signal produced at the cathode has the same polarity as that at the grid. A positive-going change on the grid causes an increase of anode current through the valve and the cathode moves more positive to chassis; i.e. the cathode 'follows' the direction of the voltage change applied to the grid. Thus a negative-going video signal of perhaps 60 volts amplitude gives rise to an identical negative-going signal of only a little less than 60 volts amplitude, at the cathode.

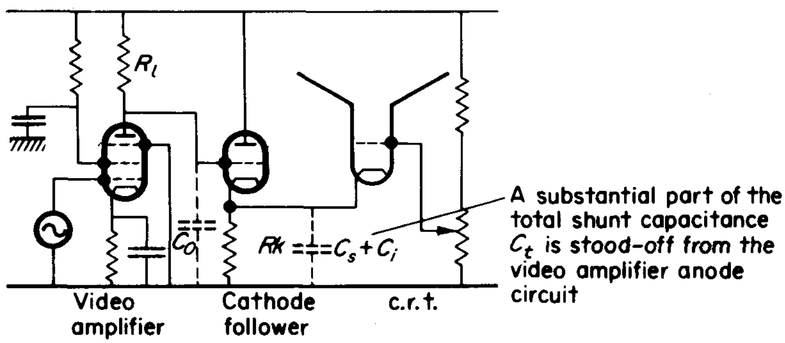



Fig. 6.8 Basic cathode follower circuit arrangement

(b) and (c) *High input impedance.* The very high input impedance effectively removes all loading from the video amplifier anode circuit. The input capacitance to a cathode follower is approximately equal to $C_{ga}$ (i.e. to the anode-to-grid capacitance of the valve).* This is of the order of 1·5 pF for a typical triode valve used in this position in a T.V. receiver.

The following example serves to illustrate how the removal of shunt capacitance from its anode circuit can affect the performance of a video amplifier.

Using some of the figures computed earlier, let $R_l = 2{\cdot}5$ kΩ, $C_t = 20$ pF and $g_m = 6$ mA/volt. This results in a medium frequency gain of 15 and a value for $f_2$ of 3·2 Mc/s. (See page 97.)

Suppose that the insertion of a cathode follower between the video amplifier and the tube removes 15 pF of the 20 pF capacitance shunting $R_l$, i.e. $C_t$ now becomes 5 pF. Several possibilities may be considered:

(i) *If $R_l$ remains at* 2·5 kΩ

Since $C_t$ has been reduced by a factor of $\frac{1}{4}$, then $f_2$ increases by a factor of four times to $4 \times 3{\cdot}2 = 12{\cdot}8$ Mc/s. The gain remains at 15 as before.

(ii) *If $R_l = 5$ kΩ*

Compared with (i) $R_l$ is now increased by a factor of 2, and the bandwidth is therefore reduced from 12·8 Mc/s to 6·4 Mc/s whilst the gain increases to $2 \times 15 = 30$.

* Note that the corresponding input capacitance for a normal resistive loaded (grounded cathode) amplifier is given by the expression

$$C_i = C_{gc} + C_{ga}(1+M)$$

where $M$ is the stage gain and $C_{gc}$ is the grid-to-cathode capacitance.

(iii) *If* $R_l = 10$ kΩ

Since $R_l$ is increased by a factor of 4 times, the bandwidth reduces once again to 3·2 Mc/s and the gain increases to $4 \times 15 = 60$.

In this hypothetical case, therefore, the insertion of a C.F. provides the following possible performance changes:

(1) Four times the bandwidth at the same gain as before.
(2) Twice the bandwidth with twice the gain.
(3) The same bandwidth with four times the gain.

It must be well understood that these particular changes are related only to the specific example cited. The ratios of gain and bandwidth changes result from the premise that $C_t$ has been reduced by the numerically convenient factor of $\frac{1}{4}$. However, the values of capacitance quoted are not too unrealistic, and the example serves to show that a cathode follower can have substantial effects upon performance.

(d) *Output impedance.* It has been tacitly assumed that the presence across the C.F. cathode resistor of the 15 pF capacitance removed from the anode load resistor will not cause a fall off in gain at the H.F. end of the range. That this assumption is justified may be checked by comparing the reactance at 3 Mc/s of this 15 pF capacitance with the output resistance of the cathode follower. In a typical case the latter may be 200 ohms.

At 3 Mc/s the reactance $X_c$ of a 15 pF capacitor is:

$$\frac{1}{\omega C} = \frac{1}{2.\pi.3 \times 10^6 \times 15 \times 10^{-12}} \text{ ohms}$$

$$= \frac{10^6}{90\pi}\Omega = 3540\ \Omega$$

Similarly, at 6 Mc/s the reactance of a 15 pF capacitor is $\frac{1}{2} \times 3540 = 1770\ \Omega$. It is evident that the shunting effect of the capacitor does not cause a significant drop in the effective output impedance of the cathode follower, and thus the performance of this additional stage does not in itself cause a drop in the overall response at high frequencies.

Before leaving the subject of C.F. output impedance it is worth noting that the figure of 200 ohms mentioned is actually the effective a.c. slope resistance of the valve.

The effective a.c. slope resistance of a valve when working as a cathode follower is given by:

$$r_a' = \frac{r_a}{1+\mu} = \frac{r_a}{1+g_m r_a} \simeq \frac{r_a}{g_m r_a} \simeq \frac{1}{g_m} \text{ ohms*}$$

To be strictly correct, the output impedance of the stage is actually this a.c. slope resistance in parallel with the cathode resistor. However, in most applications, the resistor fitted is so much higher in value than $1/g_m$ ohms that the net a.c. output resistance is only marginally less than $1/g_m$ ohms.

In the circuit shown in Fig. 6.9, a PCF80 valve is employed, with the R.F. pentode section working as video amplifier and the triode as a cathode follower.

The amplifier was designed to work in a 405/625 dual standards receiver. On the 625-line signal a bandwidth of some 5·5 Mc/s is required. Though only 3 Mc/s is needed on 405, there is

* In the formula $r_a' = 1/g_m$ ohms, $g_m$ must of course be expressed in fundamental units; i.e. $g_m$ must be in amps-per-volt.

obviously no disadvantage in having a wider bandwidth than is really necessary when receiving this signal. The only possible source of trouble is that the inter-I.F. beat frequency of 3·5 Mc/s falls well within the bandwidth, but a 3·5 Mc/s rejector circuit is switched into the pentode cathode on 405 to remove this interference. As explained in Chapter 5 on video detectors, the presence of such a tuned circuit in the video amplifier cathode, causes a high level of negative feedback to be set up at the inter-I.F. beat frequency. Should a component at this frequency be delivered from the detector to the grid input circuit, it suffers cancellation because an opposing voltage is set up across the cathode tuned circuit.

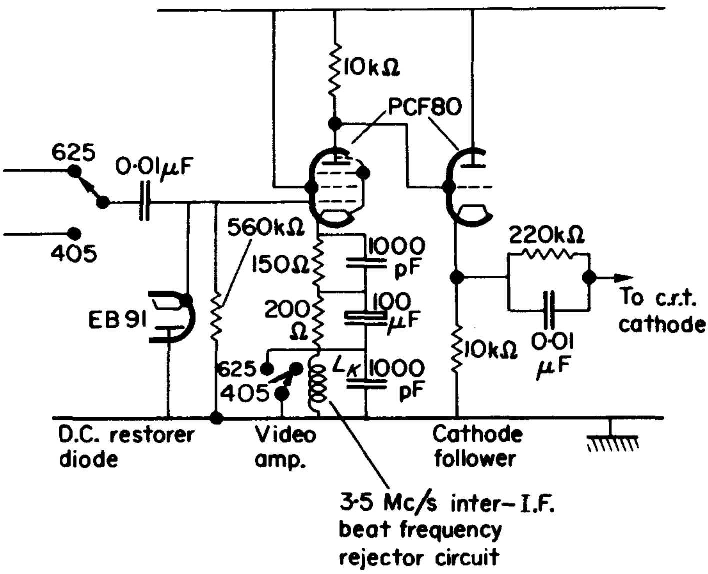



Fig. 6.9 Dual-standards receiver, video-amplifier and cathode-follower circuit

If the grid of a C.F. valve is direct coupled to the anode of the previous stage, the steady no-signal voltage developed at its cathode will be almost exactly equal to that present at the previous anode. It is obvious that the cathode cannot become much more positive than the grid since this would imply that a negative bias were being developed between grid and cathode. This would tend to decrease the anode current and so reduce the cathode voltage. On the other hand, if the grid were positive to the cathode, grid current and heavy anode current would flow, which in turn would increase the cathode voltage to remove the net positive voltage on the grid.

The natural automatic tendency of the circuit is therefore to maintain the net grid-cathode voltage, at or near zero.

Clearly, a cathode resistor must be chosen which has a value such that the appropriate voltage is dropped across it by the normal level of anode current recommended for the C.F. valve. It is often the case that the C.F. cathode resistor is rather larger in value than the anode load of the previous stage; the discrepancy being accounted for by the different current ratings of the two valves. In the circuit of Fig. 6.9, however, the two resistors are both of 10 kΩ.

A point of disadvantage of the C.F., sometimes overlooked, is that should the valve emission fail, the cathode voltage disappears. If, as is often the case, cathode modulation of the c.r.t. is

used, the c.r.t. cathode is carried downwards towards chassis potential. Since the c.r.t. grid is of course connected to a positive potential to set up the correct brightness bias voltage, the result of this fault is that the tube has a high positive voltage between grid and cathode which causes excess beam current to flow. Protection against this eventuality may be achieved by returning the tube cathode to H.T.+ via a separate resistor. This ensures a positive potential on the tube cathode even when the C.F. is inoperative.

When the tube cathode is fed from the video amplifier anode, however, the opposite situation results. Here failure of the video amplifier causes the anode voltage to rise to the full H.T. voltage, and this lifts the c.r.t. cathode voltage to a more positive level and the beam current is cut off altogether. The point will be appreciated more clearly when methods of feeding the c.r.t. are discussed.

*Method 2. The shunt peaking coil* (Fig. 6.10)

Here a small inductor is added in series with the anode load resistor $R_l$. Typically such a coil will have an inductance of the order of 50 μH to 250 μH. Though fitted in series with $R_l$, the coil is called a shunt peaking coil because it is in fact a part of the shunt anode circuit. This is illustrated in the equivalent circuit of Fig. 6.10(b).

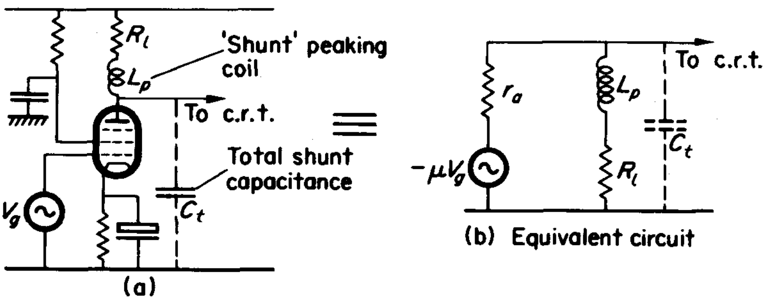



Fig. 6.10 The shunt peaking coil

The effective circuit in shunt with the signal path now consists of $C_t$ in parallel with the series combination of $R_l$ and $L_p$.

The presence of the inductor in series with $R_l$ gives a lift to the net anode circuit impedance at the H.F. end of the frequency range being handled. This partly compensates for the decreasing reactance of $C_t$.

A design formula frequently used to establish the correct order of value for $L_p$ is:

$$L_p = n.R_l^2C_t\,*$$

(where $n$ is a numerical factor whose value is usually of the order $n=0{\cdot}25$ to $n=0{\cdot}7$).

The curves of Fig. 6.11 illustrate the effect upon an amplifier's response curve of the insertion of shunt peaking coils of various values.

Instead of plotting specific curves for a particular amplifier with a given value of $R_l$ and $C_t$, the curves are 'generalised' so that they are applicable to *any* amplifier. This is done by plotting the ratio (Gain at H.F.)/(Gain at M.F.) on the $y$-axis against the ratio $f/f_2$ on the $x$-axis for

* See Appendix.

various values of the factor $n$. Note that $f$ is the test frequency and $f_2$ the $-3$ dB frequency of the uncompensated amplifier.*

When $n=0$, $L_p=0$, so that the curve labelled $n=0$ shows the response of an amplifier without compensation.

This particular curve shows a drop of relative gain of 0·707 (i.e. $-3$ dB) when the frequency $f$ fed into the amplifier equals $f_2$ (i.e. when the ratio $f/f_2=1$). This is obvious since by definition $f_2$ is the frequency at which the gain of the uncompensated amplifier falls by 3 dB.

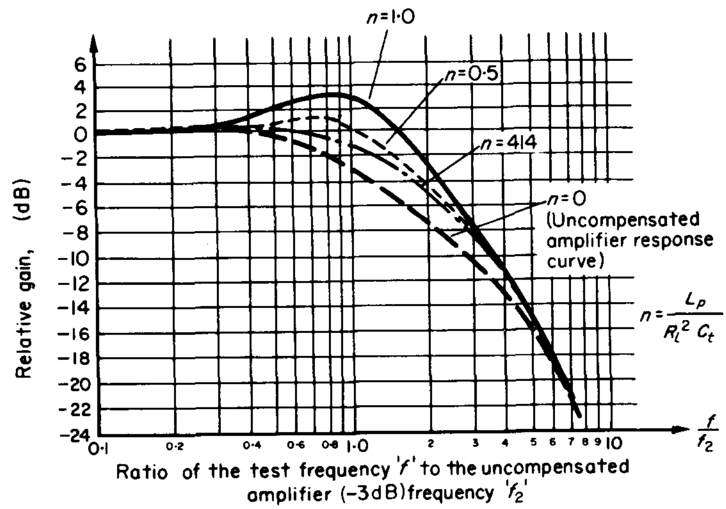



Fig. 6.11 Video amplifier gain/frequency response curves showing the effect of shunt-peaking coils of various inductances

It should also be noted that the $y$-axis is calibrated in decibels.

The expression (Gain at H.F.)/(Gain at M.F.) refers to a *voltage* gain ratio, and to translate a voltage ratio into decibels it is necessary to apply the formula:

$$\text{Relative gain in dB} = 20 \log_{10} \text{(voltage ratio)}$$

When the 'Gain at H.F.' equals the 'Gain at M.F.' the voltage gain ratio is unity, hence:

$$\text{Relative gain in dB} = 20 \log_{10} \text{(voltage gain ratio)} = 20 \log_{10} 1 = 0 \text{ dB}$$

The 0 dB line hence forms the reference level on the gain/frequency response curve.

When the gain at H.F. falls below that at M.F., the expression yields a negative number of decibels, and conversely when the gain at H.F. is greater than the gain at M.F., the ratio shows a positive number of decibels.

Referring once again to Fig. 6.11, as $n$ is increased (i.e. as the inductance of the peaking coil increases), the response curve lifts at the H.F. end. The curves show that when $n=0{\cdot}5$ there is complete 'equalisation' of the response curve at the half-power frequency $f_2$ (i.e. at the point

*For a particular amplifier the $x$-axis may readily be converted to read frequency directly if desired; e.g. suppose in an uncompensated amplifier the gain falls by 3 dB at a frequency of 3 Mc/s (i.e. $f_2=3$ Mc/s). Then, at the point $f/f_2=1$, $f=f_2=3$ Mc/s; the point $f/f_2=2{\cdot}0$ corresponds to 6 Mc/s; $f/f_2=3{\cdot}0$ is 9 Mc/s, etc.

In another case, if $f_2=4{\cdot}5$ Mc/s, then the point 1·0 on the $x$-axis corresponds to 4·5 Mc/s; the point 2·0 is 9 Mc/s, etc. The examples illustrate the universal nature of the curves.

$f/f_2 = 1$, the gain is exactly equal to the gain at M.F.). That this is so is demonstrated in the Appendix at the end of the book. The value $n = 0{\cdot}5$ is a popular approximate value for $n$ in practical circuits.

It will be noticed that there is a small hump just below this frequency. Increasing the inductance increases the amplitude of the hump and also steepens the rate at which the gain falls off above $f_2$. It is characteristic of compensating coils that, whilst they lift the response curve in the desired region, the subsequent fall-off of gain is more rapid than in the uncompensated circuits. Too steep an edge to the response curve is not a desirable characteristic since it can give rise to a tendency to produce 'overshoot' on transients. The matter is discussed later in the chapter.

At this point a common source of misconception must be mentioned. Reference to Fig. 6.10 shows that the anode circuit comprising $L_p$, $R_l$ and $C_t$ takes on the appearance of a parallel-tuned circuit, and the hump which appears in the frequency response curve is sometimes wrongly assumed to be due to resonance in this tuned circuit. In most practical circuits, however, the value of inductance employed is so small, and the ohmic value of $R_l$ is so large (i.e. large in relation to the inductance), that the circuit is in fact completely non-resonant. It is demonstrated in the Appendix that when the factor $n$ is equal to, or less than, unity, the circuit is non-resonant. As stated below, $n$ normally has a value of less than unity in video amplifiers.

In the practical development of a particular circuit it is usual to start off with an inductor of the order of value indicated by the formula $L_p = 0{\cdot}5\ R_l^2 C_t$ and then to experiment with larger or smaller inductors until the desired response is obtained. The value of $R_l$ is presupposed to have been chosen to make $f_2$ approximate to the highest frequency to be handled.

It should be noted that ideally the following three characteristics are desirable:

(1) Constant gain throughout the pass-band.
(2) Linear phase response (i.e. phase shift should be proportional to frequency, for the reasons discussed earlier).
(3) Maximum transient response without overshoot.* This is often defined as the condition of 'critical damping'.

It is possible to deduce what value $n$ must have in the formula $L_p = nR_l^2 C_t$ to satisfy each of these requirements in turn. Unfortunately, each leads to a different answer!

Thus:

(1) For optimum frequency response: $n = 0{\cdot}414$
(2) For optimum phase response: $n = 0{\cdot}322$
(3) For critical damping: $n = 0{\cdot}25$

In practice, a value of $n = 0{\cdot}5$ is often chosen which gives equalisation at $f_2$; a small hump below $f_2$; and a slight but tolerable (perhaps beneficial) tendency towards overshoot on transients.

*Example.* It was shown earlier that in an amplifier using a pentode of $g_m = 6$ mA/V, with an anode load resistor $R_l$ of 2·5 kΩ and a total shunt capacitance $C_t$ of 20 pF, the gain at mid frequencies is 15 and the $-3$ dB frequency $f_2$ is 3·2 Mc/s. (See page 97.)

Suppose it is decided to fit a shunt peaking coil using a value of $n = 0{\cdot}5$, in order to lift the gain at $f_2$ by 3 dB; i.e. to give equalisation at 3·2 Mc/s.

* A transient is a steep-fronted waveform of the kind produced by a sharp edge (say black to white) in picture detail. Overshoot gives rise to the effect known as 'ringing' on the reproduced picture.

Then
$$\begin{aligned} L_p &= nR_l^2C_t \\ &= 0{\cdot}5 \times 2500^2 \times 20 \times 10^{-12}\ \text{H} \\ &= 0{\cdot}5 \times 2500^2 \times 20 \times 10^{-12} \times 10^6\ \mu\text{H} \\ &= 0{\cdot}5 \times 2{\cdot}5^2 \times 10^6 \times 20 \times 10^{-12} \times 10^6\ \mu\text{H} \\ &= 62{\cdot}5\ \mu\text{H} \end{aligned}$$

It is instructive to note by reference to Fig. 6.11 that the $-3$ dB frequency is shifted up to between 5 Mc/s and 6 Mc/s if an inductor of this value is fitted. (Note that the point $f/f_2 = 1{\cdot}0$ is 3·2 Mc/s and $f/f_2 = 2{\cdot}0$ is 6·4 Mc/s in this particular case.)

*Method 3. The series peaking coil*

As an alternative arrangement the peaking coil may be included in series with the signal path, rather than in shunt with it. Fig. 6.12(a) shows the position of the coil in a basic circuit and Fig. 6.12(b) is the equivalent a.c. circuit.

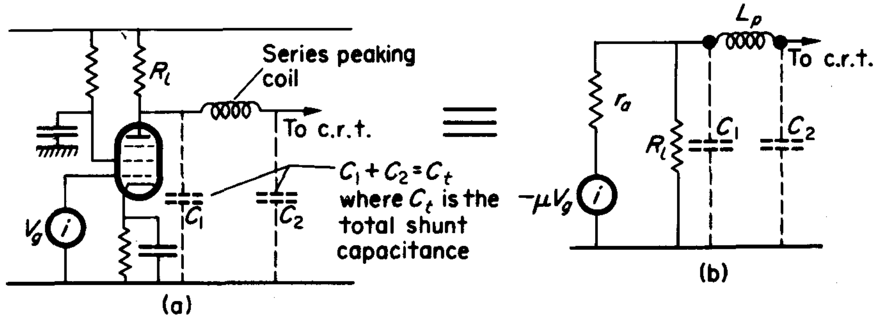



Fig. 6.12 The series peaking coil

In practice the coil is usually fitted very close to the anode pin of the valve and in this position it effectively separates the total shunt capacitance $C_t$ into two parts with $C_o$ on one side and $(C_s + C_i)$ on the other. The circuit so formed takes on the appearance of a low-pass filter, and it is as such that the principle of the series peaking coil is best studied.

The value of $L_p$ is so chosen that the filter passes all frequencies within the required video signal bandwidth with minimum attenuation, but presents a rising attenuation above the upper limit of this frequency band. In filter technology extremely accurate and carefully tailored attenuation/frequency characteristics may be obtained. Clearly, in a domestic television receiver, a simple compensation circuit wired into a video amplifier cannot be treated as a sophisticated filter. None the less, by reference to filter design theory, it is possible to choose a series inductor which, in conjunction with the circuit capacitances, behaves sufficiently well as a low-pass filter to give a substantial improvement in the H.F. response.

A useful basic design formula for the inductor is again given by:

$$L_p = n \,.\, R_l^2 C_t$$

As before, $n$ is a numerical factor, which in this case is usually ascribed a value between 0·5 and unity. Since the object is to increase the frequency at which the gain at H.F. falls by 3 dB on that at M.F., it is again convenient to use as a reference frequency the frequency $f_2$ at which the gain of the uncompensated amplifier falls by 3 dB. A family of curves may be plotted as in

Fig. 6.13 showing the relative response with various values of the factor $n$. Substitution of particular values for $R_l$ and $C_t$ allows $f_2$ to be calculated; i.e.

$$f_2 = \frac{1}{2\pi R_l C_t}$$

for a given amplifier, and by reference to the curves a value of $n$ may be chosen which gives the desired frequency response. The required value of inductance is then found by application

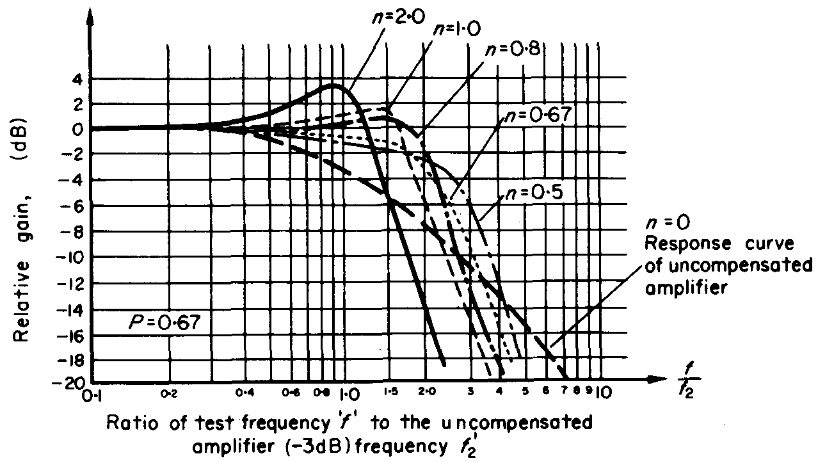



Fig. 6.13 Approximate general form of video amplifier gain/frequency response curves, showing the effect of series peaking coils of various inductances. Here the distribution of the shunt capacitance is held constant. ($C_1 = 0{\cdot}33$, $C_2 = 0{\cdot}67$, $C_t = C_1 + C_2 = 1{\cdot}0$, and $P = C_2/C_t$.)

Based upon curves published in the BBC Engineering Training Manual: *Television Engineering*, Vol. 11, by S. W. Amos and D. C. Birkinshaw; published by Iliffe & Sons Ltd., 1956. By kind permission of the BBC.

of the above formula for $L_p$. It will be noticed in Fig. 6.13 that the curves are shown as being applicable to a ratio of

$$P = \frac{C_2}{C_1 + C_2} = \frac{2}{3} \quad \text{(i.e. } C_2 = 2C_1\text{)}$$

It is obvious that the behaviour of the filter must change as the disposition of the total capacitance $C_t$ changes. Just how much appears at the input to the filter (shown as $C_1$) and at the output ($C_2$) must clearly vary according to the nature of the amplifier and its coupling circuitry. Thus a set of universal response curves for various values of $n$ and a single fixed ratio of $C_2/C_t$ is of use only to this one specific case. There is a need for a second family of curves which show the effect of varying the capacitance ratio whilst $n$ (hence the value of inductor) is held constant. For a series peaking coil a value for $n$ of 0·67 is a popular value since it gives optimum frequency response when the capacitance at the input (anode) end of the filter is $\frac{1}{3}$ as great as that at the output end (i.e. $C_2/C_t = 0{\cdot}75$). This is a typical example of how the capacitance splits up in a practical circuit. Fig. 6.14 shows a sketch of curves of this kind.

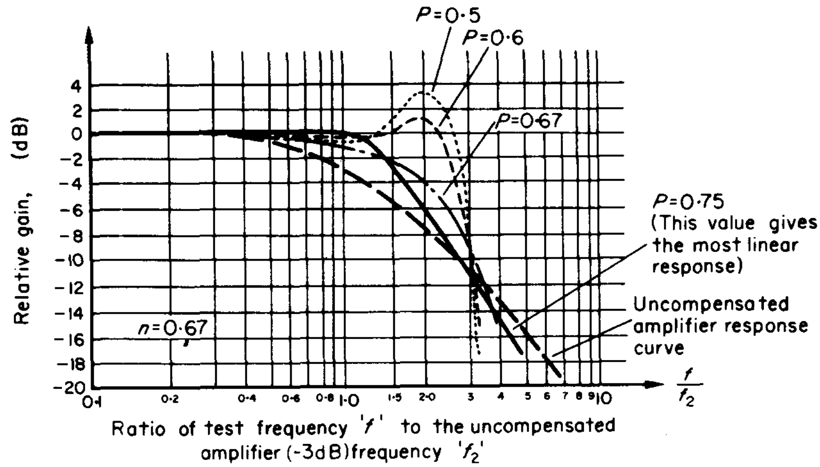



Fig. 6.14 Approximate general form of video amplifier gain/frequency response curves showing the effect of using a series peaking coil of fixed inductance and varying the distribution either side of it of the total shunt capacitance $C_t$.

$$\left(P = \frac{C_2}{C_t},\ C_t = C_1 + C_2,\ \text{and}\ n = \frac{L_p}{R_l^2 C_t} = 0{\cdot}67\right)$$

Based upon curves published in the BBC Engineering Training Manual: *Television Engineering*, Vol. 11, by S. W. Amos and D. C. Birkinshaw; published by Iliffe & Sons Ltd., 1956. By kind permission of the BBC.

It will thus be observed that there are two variables to consider in the case of the series-peaking coil:

(1) $n$, which controls the value of $L_p$ (from $L_p = nR_l^2C_t$).

(2) The factor $P = C_2/C_t$, which refers to the distribution of the total shunt capacitance $C_t$.

Clearly, the factor $P$ may be manipulated in circuit design to produce a given desirable distribution of $C_t$.

Fig. 6.15 shows an alternative form of the series peaking coil connection. Here the connection to the c.r.t. is taken from the junction of $L_p$ and $R_l$ instead of from the anode. The equivalent circuit shows that the only difference between the two circuits is that $R_l$ appears at

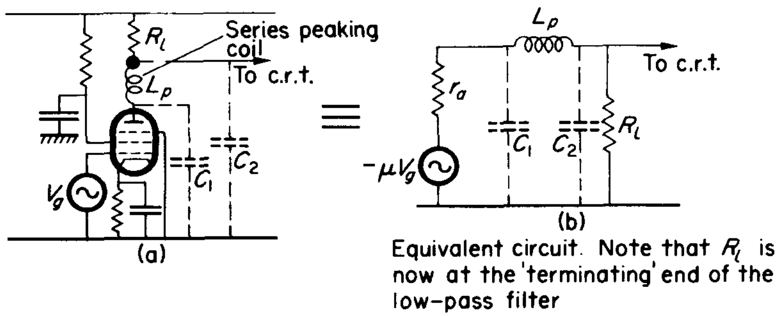



Fig. 6.15 Alternative series-peaking coil arrangement

*N.B.* The circuit is often confused with the 'shunt' peaking coil arrangement because $L_p$ and $R_l$ appear in the same positions in the circuit diagram. Notice, however, that in the series coil case the video signal is 'taken-off' from the junction of $L_p$ and $R_l$ instead of from the anode. This places $L_p$ in series with the signal path instead of in shunt with it.

the 'sending-end' of the low-pass filter in the former circuit, but at the 'receiving-end' in the latter. The same design curves cover both cases, but the capacitance ratio attached to the curves now refers to $C_1/C_t$ instead of $C_2/C_t$. Thus the numerator in the capacitance ratio refers always to the capacitance on the opposite end to that terminated by $R_l$.

It is evident from what has been said that the theory of the series peaking coil is more complex than that of the shunt coil. In practice, the best procedure to adopt, should it be necessary to deduce a suitable value for a series peaking coil in a particular circuit, is to estimate an approximate value by taking into account the factors mentioned, and then to experiment with inductors around this order of value until an acceptable response curve is produced. This is best done using a visual display system. The output from an F.M. generator which sweeps in frequency right through the *required* video range, is fed into the amplifier. The rectified output from the amplifier is presented to the 'Y' plates of an oscilloscope and a *marker generator* allows for calibration of the reproduced response curve. In addition the transient response may be checked by feeding in a square wave signal and checking the rise-time of the output waveform.

The following summarises the relative properties of the series coil as compared with the shunt coil:

(a) Part of the shunt capacitance is isolated from the video amplifier anode circuit. (Advantage.)
(b) A better rise-time performance is possible. (Advantage.)
(c) The fall-off in gain (i.e. attenuation) just beyond the required upper edge of the bandwidth is steeper. (May be a disadvantage.)
(d) The circuit is more critical to adjust for optimum performance than the shunt peaking coil arrangement. (Disadvantage, but one which affects development work only.)

*Method 4. Combined shunt and series peaking coils*

Sometimes both shunt and series peaking coils are used in the anode circuit of a video amplifier. Fig. 6.16 shows an example of such a circuit.

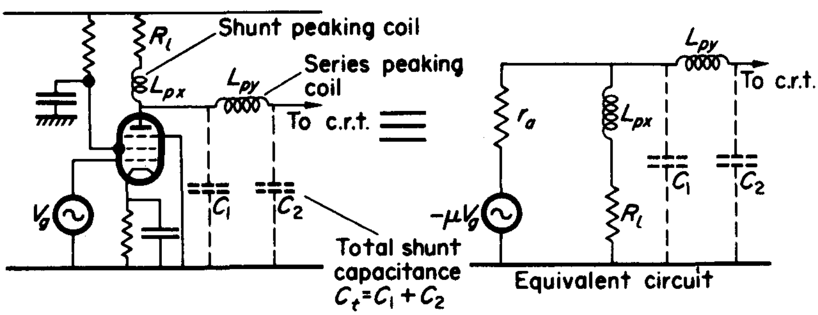



Fig. 6.16 Use of combined shunt- and series-peaking coils

The frequency response obtainable with this arrangement can be very good if the values are carefully chosen. Unless the disposition of shunt capacitance is known in advance, it is not possible to generalise over inductor values. However, if $L_{px}$ and $L_{py}$ are the shunt and series inductances, respectively, the following formulae may be used as a guide to establish the approximate order of values needed; the final choice being established by experimentation.

$$L_{px} = n_x . R_l^2 C_t \qquad L_{py} = n_y R_l^2 C_t \qquad P = \frac{C_2}{C_t}$$

As before, the reference frequency is the frequency $f_2$ at which the gain of the uncompensated amplifier falls by 3 dB; that is,

$$f_2 = \frac{1}{2\pi R_l C_t} \text{ c/s}$$

For given values of $R_l$ and $C_t$, there are thus three variables to determine. These are the disposition either side of the series inductor of the total shunt capacitance $C_t$, and the values of the two factors $n_x$ and $n_y$. As a basis to start from, the approximate values of these variables for three important criteria are tabulated below:

**Table 6.4**

*Suggested approximate values of the variables P, $n_x$ and $n_y$ for various performance characteristics (see Fig. 6.17)*

| Criterion | $P$ | $n_x$ | $n_y$ |
|---|---|---|---|
| (1) Linear frequency response | 0·6 | 0·14 | 0·58 |
| (2) Optimum phase response | 0·72 | 0·1 | 0·46 |
| (3) Critical damping | 0·8 | 0·063 | 0·39 |

The sketch graphs of Fig. 6.17 show the approximate form which the gain/frequency response curve takes in these three cases. Should peaks occur in the response curve due to self-resonance of the peaking coils, shunt damping resistors are fitted. This is particularly true of the series inductor, and a suitable value of resistor for connection in parallel with $L_{py}$ is of the order of $2 \times R_l$, where $R_l$ is the anode load resistor.

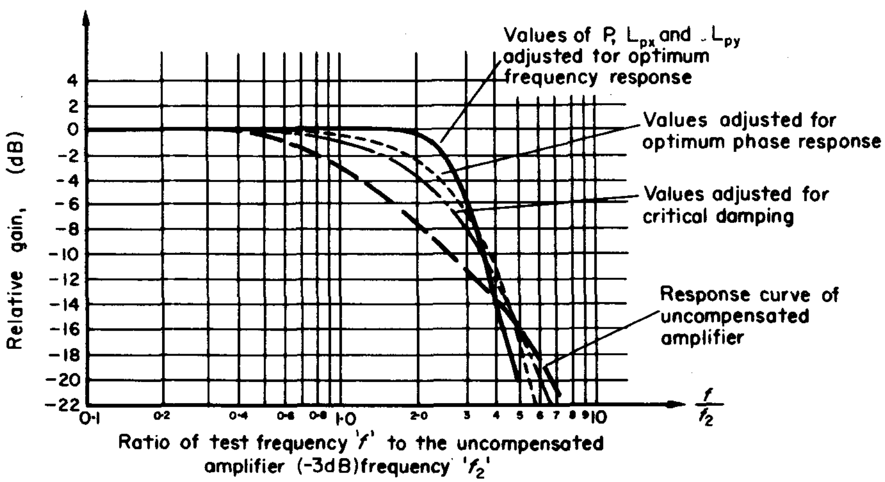



Fig. 6.17 Gain/frequency response curves of a video amplifier employing shunt-series compensation (see Table 6.4)

After original source in *Television Engineering Handbook*, edited by D. G. Fink, McGraw-Hill, 1957. Used by permission of McGraw-Hill Book Company.

The shunt-series peaking coil arrangement is capable of yielding a rather better result than is possible by the use of either of the coils individually.

It should be mentioned that a filter circuit such as the low-pass filter to which the series coil approximates, only gives correct results when it is correctly terminated. The basic circuit is designed to match the anode load resistor $R_l$ in the middle-frequency region. However, as the frequency rises towards the upper end of the range handled, the impedance necessary to give correct termination (i.e. matching), rises quite sharply above $R_l$ and the degree of mismatch, hence loss, increases accordingly. When a shunt coil exists in series with $R_l$, the series impedance of these two components now represents the termination 'seen' by the low-pass filter. In the M.F. region, where the shunt inductor has only small reactance, the filter is still terminated correctly by $R_l$. At the H.F. end, the rising reactance of the shunt coil adds to the net terminating impedance with the result that a somewhat better match is achieved than before.

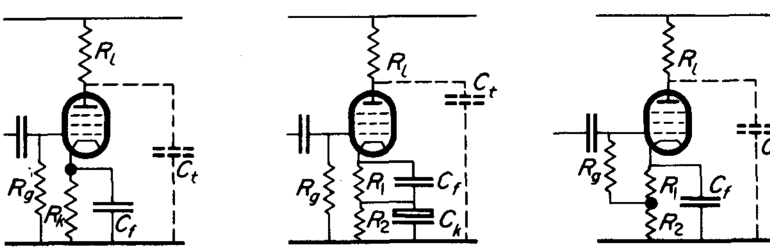
*Method 5. Cathode compensation*

The basic principle of this method is to apply negative feedback over the lower and middle frequency regions, but to arrange for it to be progressively removed in the H.F. region. This is a case where some overall gain is sacrificed in order to arrive at a more even response curve.

The circuit is very simple and merely involves the selection of a cathode by-pass capacitor of value such that it fails to decouple the cathode feedback resistor over the main part of the frequency range. At higher frequencies the decreasing reactance of the capacitor reduces the net impedance of the C.R. cathode circuit and the negative feedback voltage decreases. The overall effect therefore is that the fall-off in output caused by $C_t$ in the uncompensated amplifier is counteracted by the increasing amplifier gain which results from the progressive removal of negative feedback.

Of course it is unlikely that the value of resistor needed for cathode bias purposes is also the optimum value for feedback purposes.

The diagrams of Fig. 6.18 show suitable arrangements when the feed-back resistor is equal to, less than, or greater than the required bias resistor. In all cases $C_f$ is the compensating capacitor. Where the total bias resistance is larger than the required feedback resistance, the unwanted portion of the bias resistance is heavily decoupled by a large capacitance. Where the bias




(a) Bias and negative feedback resistors are equal. (b) Bias resistance is higher than feedback resistance. (c) Feedback resistance is higher than bias resistance.

Fig. 6.18 Cathode compensation arrangements

(a) $R_k$ serves to provide both the required bias and the negative feedback at L.F. and M.F.

(b) $(R_1+R_2)$ provide bias, but only $R_1$ serves for feedback purposes.

(c) $R_1$ provides bias, but negative feedback is given by $(R_1+R_2)$.

*N.B.* In all cases the value of $C_f$ is so chosen that it progressively 'removes feedback' at the H.F. end of the frequency range.

resistor needs to be smaller than the feedback resistor, the grid is returned to the junction of two cathode resistors, with the upper one forming the bias resistor.

On occasions, variable capacitors are fitted in the position of $C_f$ and these allow adjustment of the H.F. response. If the gain at H.F. is relatively too high, it is possible for 'overshoot' to take place on steep-fronted waveforms. This may give rise to the well-known symptom of 'ringing' on the picture and the capacitor should then be adjusted to the point where ringing just fails to be apparent.

Yet another modification is the insertion of an inductor in series with $C_f$. This forms a series tuned circuit which is made to resonate near the upper edge of the video frequency band. At resonance the very low impedance of the series tuned circuit effectively short circuits the feedback resistor and the negative feedback is virtually reduced to zero.

As a guide to the values of components needed, the time constant ($R_k C_f$) of the cathode circuit should be of the same order as the time constant ($R_l C_t$) of the anode load resistor and shunt capacitance. Typically the ratio of cathode circuit to anode circuit time constants falls within the range 0·5 to 2·0.

**Compensation at the L.F. end**

With wideband amplifiers for general purposes it is sometimes necessary to compensate for the L.F. droop by including circuitry which gives a lift to the gain at this end of the range. With R.C. coupled amplifiers the lower half-power frequency is given by:

$$f_1 = \frac{1}{2\pi C_c R_g}$$

Clearly, $f_1$ may be lowered by increasing $C_c$ and $R_g$. If direct coupling is used, $C_c$ is dispensed with and the problem is then largely solved. However, decoupling capacitors in the screen and cathode of a pentode cease to be as efficient as the frequency gets lower and the result is the appearance of increasing negative feedback at the L.F. end of the range. The simple device illustrated in Fig. 6.19 may be made to obviate this effect.

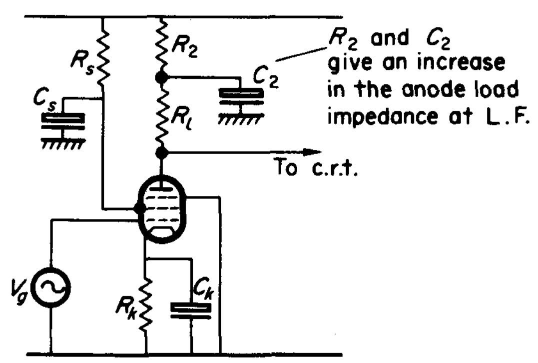



Fig. 6.19 L.F. compensation circuit

Compensation is achieved by the decoupling components $R_2$ and $C_2$ which are inserted in the anode circuit. At H.F. and M.F. the reactance of $C_2$ is so low that $R_2$ is short circuited and the 'top-end' of $R_l$ is grounded in the normal way. As the frequency falls to a low level,

however, the reactance of $C_2$ rises and the effective anode load now consists of $R_l$ in series with the parallel impedance of $C_2$ and $R_2$. At the L.F. end, therefore, to compensate for the tendency for the gain to decrease, due to the effect of increasing negative feedback, the net anode load impedance of the stage increases, so that the actual final gain achieved remains more nearly constant.

With video amplifiers in television receivers it is often found that rather than there being efforts made to enhance the L.F. response, there are, on the contrary, measures taken to degrade it. This is often done to reduce the flutter or 'pumping' effect caused by aircraft reflections. As an aircraft approaches there is a continuous change of phase between the direct signal and the one reaching the receiver by reflection from the aircraft. The two signals beat together and the constantly changing phase relationship causes amplitude modulation of the resultant arriving signal. As the aircraft gets nearer, the beat frequency changes and at the same time the reflected signal gets stronger. This results in an increasingly deep rise and fall of the net resultant signal reaching the receiver. Although efficient fast-acting a.g.c. systems are capable of reducing this effect, it is usual to include a long time constant circuit in series with the video signal path to the tube cathode, to make the result less annoying.

Fig. 6.20 shows a typical circuit arrangement.

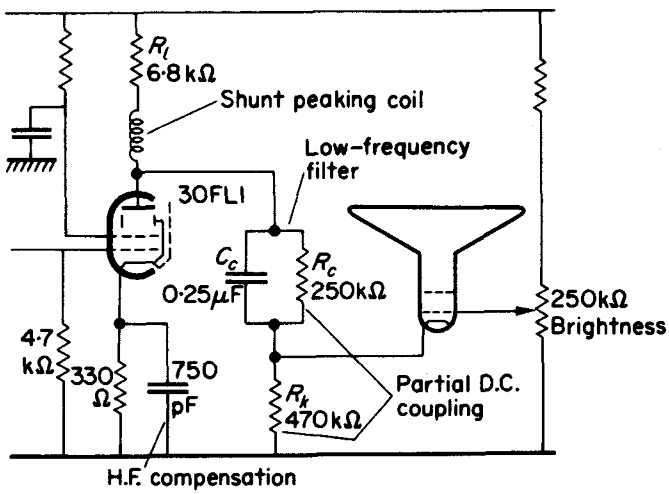



Fig. 6.20 Example of a circuit employing a low-frequency filter and partial d.c. coupling between the video amplifier and the c.r.t.

$C_c$ has a low reactance at M.F. and H.F., but at very low frequencies it ceases to by-pass $R_c$ with the result that this resistor provides series attenuation in the signal path to the offending low frequency pulsations. Table 6.5 gives examples of typical component values found in this circuit. It is worth noting that if $C_c$ is open circuited, the M.F. and H.F. response is so severely attenuated that the picture becomes blurred, 'plastic' and totally lacking in detail.

If the resistor $R_c$ were not present, there would be a complete loss of the d.c. component. With $R_c$ connected the series attenuation to the d.c. component is $R_k/(R_c+R_k)$. This is sometimes

**Table 6.5**

*Examples of parallel C.R. networks found in series with the video signal path to the c.r.t.*

| $C_c$ | $R_c$ | Time constant |
|---|---|---|
| (1) 0·01 μF | 220 kΩ | 2·2 ms |
| (2) 0·1 μF | 330 kΩ | 33 ms |
| (3) 0·1 μF | 470 kΩ | 47 ms |
| (4) 0·22 μF | 270 kΩ | 59·4 ms |
| (5) 0·5 μF | 220 kΩ | 110 ms |

referred to as *partial* d.c. *coupling*. Apart from the decrease of aircraft flutter, the circuit shown confers other advantages:

(a) The H.T. voltage on the tube cathode is reduced from the video amplifier anode voltage of $V_a$ to

$$V_a \times \frac{R_k}{R_c + R_k}$$

This lessens the voltage stress between the tube cathode and heater. To give maximum protection from burn-out due to faults in series heater chains, the c.r.t. heater is located at, or near the earthy end of the chain. With direct coupling between the video amplifier anode and the tube cathode, almost the full H.T. voltage is applied between cathode and heater. The possibility of heater-cathode insulation breakdown is reduced when the working voltage is brought down by the potential divider effect of $R_c$ and $R_g$.

(b) The reduction in the working voltage of the c.r.t. cathode means that both cathode and grid are nearer to the H.T. negative (i.e. chassis) potential. Therefore the net positive H.T. voltage, relative to cathode, which is available (e.g. from the 'boost-H.T.-rail') for feeding the tube's first or focus anodes, is greater by the amount of the reduction in positive cathode voltage relative to chassis.

(c) Since the d.c. component is partly removed, the difference in mean level brightness from scene to scene is reduced. This restricts the overall contrast range which has to be handled by the tube and hence limits the maximum demand which is made upon the e.h.t. system. The regulation of the normal type of e.h.t. system used in domestic receivers is not good and excessive current drain results in a falling voltage. This in turn leads to an instantaneous reduction in deflection sensitivity, so that the picture expands ('blooms') as the voltage falls. The increase in deflection sensitivity is of course due to the fact that as the e.h.t. voltage falls, the forward velocity of the electrons in the c.r.t. beam decreases. The electrons hence 'spend a longer time' in the deflecting field and are therefore deflected further, by a given field, than they would normally be. A more constant performance is produced by artificially restricting at the receiver the natural range of brightness levels which occur in the original transmitted picture. The partial loss of the d.c. component is the lesser of two evils, and involves a commercial compromise. In terms of absolute picture fidelity both *complete retention of the d.c. component* and *a well regulated e.h.t. system* are necessary.

**Transient response**

Transients are signal voltage changes of a steep-fronted 'square-wave' nature. The information along a video signal line is essentially of square-wave form. It is of course very seldom indeed that the change of brightness along a line is sinusoidal in nature. In general, the reproduction of detail along a line necessitates more or less sharp changes in the modulation voltage level. Edges require 'steps' in voltage; either up or down depending upon the relative brightness either side of the edge. Less well-defined detail requires gradually increasing or decreasing changes in the video voltage level.

As stated in Chapter 4, square waves, and indeed any complex waveforms, may be analysed mathematically and shown to be equivalent to the algebraic sum of a whole series of individual sine wave harmonics of varying discrete amplitudes. Because this is so it is possible to analyse the finest square wave detail which the system has to portray, and to deduce what frequency range has to be handled in order to give good reproduction of this detail. This leads to the frequency and phase response characteristic of video amplifiers, and indeed ultimately of the whole vision receiver, being examined entirely in terms of sine waves.

Whilst this is a useful guide and affords a yardstick by which performance may be examined, it is none the less often difficult, if not impossible, to interpret the results in terms of the effects upon the reproduced picture itself. Plots of amplifier gain against frequency, and of phase change against frequency, may perhaps both show substantial departures from the sought-after optimum shapes, but it is extremely difficult to foresee just what visible effect such conditions may produce upon a picture.

For example, suppose the picture being resolved consists in one area of alternate vertical black and white columns. The corresponding video signal along a line passing through these columns will show a regular square wave voltage change. From the time duration of each cycle the square wave frequency may be deduced, and this in turn may be analysed to show the frequencies and amplitudes of a whole series of individual sine wave harmonics which if added together reproduce the same square wave.

With a perfect idealised amplifier all of these components would pass through and be treated both to an equal degree of amplification and to a degree of phase shift proportional to their frequencies. The square wave would then appear at the output terminals, amplified but otherwise unchanged. If, however, the gain-frequency and the phase-frequency characteristics were not perfect, then the relative amplitudes and phases of all the component sine waves would be changed. Clearly, to see what effect this would have involves the laborious algebraic additions of all the components so that the final waveshape may be arrived at.

To sum up, it is evident that to endeavour to assess the performance of video equipment by plotting gain-frequency and phase-frequency characteristics leads to difficulties, since this information does not yield a direct indication of how efficiently the apparatus will reproduce picture detail.

A better method is to inject waveforms whose shape is typical of the voltage changes present in normal video signals. If these are then examined at the output terminals of the amplifier, a direct impression of distortion is available and, with practice, the nature of the distortion will give the observer an immediate assessment of just how well the system will resolve a normal picture.

This requirement has led to the use of the so-called 'pulse and bar' test signals. Such signals are designed to give a direct measure of the transient response of the system. Fig. 6.21 shows the nature of the waveforms sometimes used for testing the British television systems. The pulse

is a sudden rise and return from black level to peak-white level. This would correspond in a normal picture to a spot of bright light on a dark background and clearly its reproduction calls for a good high frequency response. The bar represents a sudden step-voltage change from black to peak white, and then requires the system to maintain this level for a comparatively long time before a return to black level is made. The edges of this bar give an indication of high-frequency response whilst the top shows information about low frequency response.

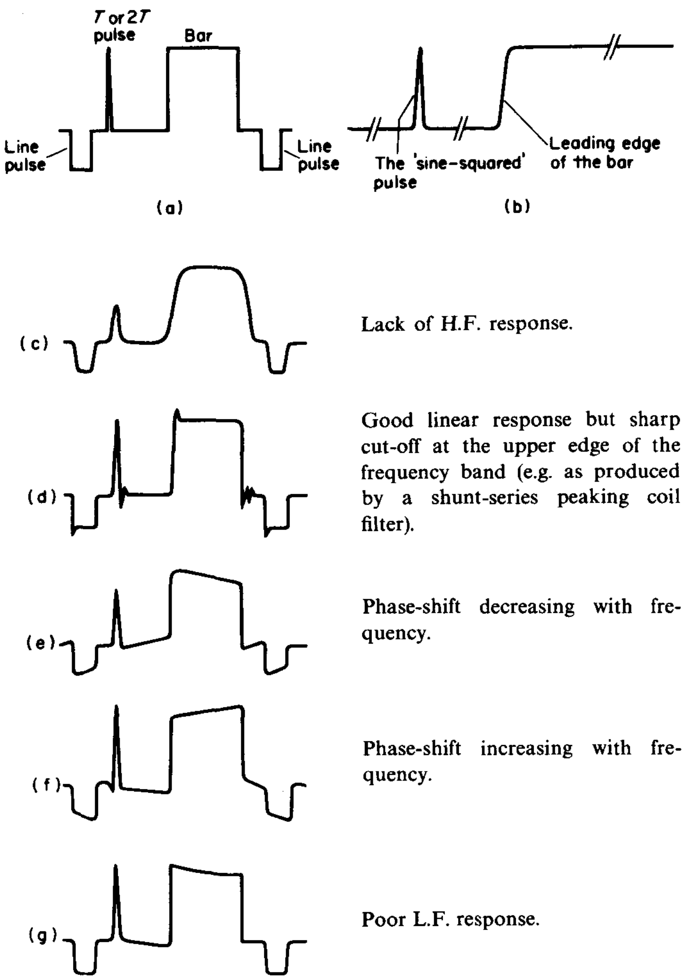



Fig. 6.21 The pulse and bar test waveform

The pulse is in fact not a square pulse but of 'sine-squared' shape. This means that it has the same shape as the square of the positive half cycle of a normal sine wave. As mentioned before, electrical circuits have a finite rise-time and by choosing a sine-squared pulse shape instead of a square (vertical-edged) pulse, the waveform (though difficult to generate) possesses the correct passband and hence allows for a realistic test of circuit performance. To choose as a test waveform, one which is beyond the capacity of normal apparatus to reproduce is obviously not the best basis on which to set up a test procedure.

To give a measure of performance over the entire frequency range of a video signal two alternative pulses are really needed. These are designated the $T$ and the $2T$ pulses. The $T$ pulse has a half-amplitude duration $T$ such that $T$ is half the periodic time of the highest video frequency used by the system. For example, if the highest video frequency is 3 Mc/s, the periodic time is $\frac{1}{3}$ μs, and $T=\frac{1}{2}\times\frac{1}{3}=\frac{1}{6}$ μs. The wider pulse has a corresponding width which is twice as great. To sum up, the bar gives an indication of performance from a few kc/s up to, say, 0·5 Mc/s; the $2T$-pulse gives information about the range 0·5 Mc/s up to the order 3 Mc/s, and the $T$-pulse extends the range up to the region of 6 Mc/s.

In the sketches of Fig. 6.21, the effect upon the test waveform of a variety of inadequate performance conditions is illustrated. The diagrams are exaggerated to show the pattern of change to be expected.

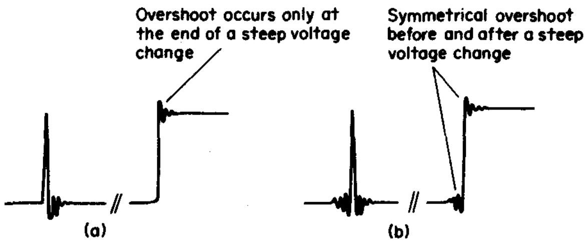



Fig. 6.22 Influence of phase response on the overshoot produced by a sharp H.F. cut-off
(a) Sharp H.F. cut-off accompanied by non-linear phase response.
(b) Sharp H.F. cut-off accompanied by linear phase-response.

It is interesting to note the influence of phase response on overshoot. A sharp H.F. cut-off, if accompanied by a non-linear phase response, shows up as overshoot at the *end* of any sharp voltage change. The same cut-off characteristic when accompanied by a linear phase response shows up as symmetrical overshoot at the beginning and end of such sharp voltage changes. The effect is illustrated in the sketches shown in Fig. 6.22.

# 7

# Video Amplifiers: Specimen Valve Circuits

In the previous chapter, general wideband amplifier techniques were studied. In the present chapter some typical valve video amplifier circuits are examined so that the practical application of the principles may be identified.

The circuit shown in Fig. 7.1 reveals the following features:

(a) An R.F. pentode of $g_m = 7{\cdot}4$ mA/V is used as the video amplifier. With an anode load resistor of 4·7 kΩ, this gives a maximum possible middle frequency gain of approximately 35 (i.e. Gain$\simeq g_m R_l$).

(b) A shunt peaking coil $L_p$ is used in the anode circuit to give H.F. compensation.

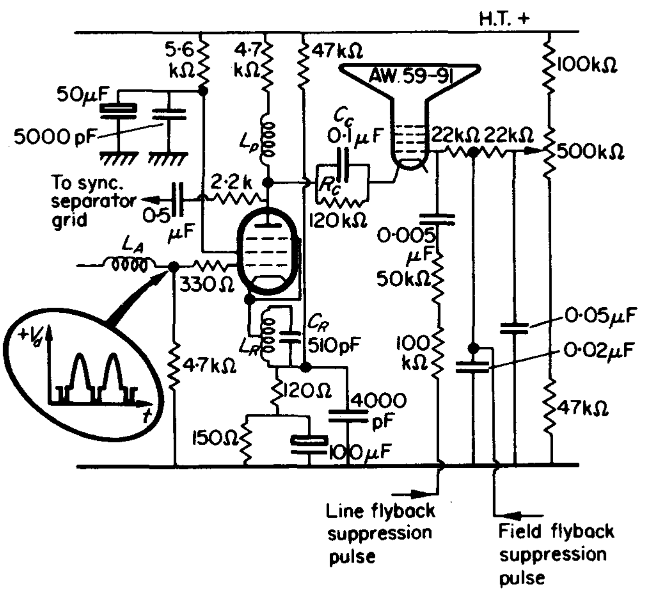



Fig. 7.1 Typical video amplifier circuitry

(c) A measure of H.F. compensation is also evident in the cathode circuit. Bias is developed partly by the valve's cathode current through the 120 Ω and 150 Ω series resistors, and partly by the bleed current passed through the resistors by the 47 kΩ resistor which returns to the H.T. line. The latter gives a measure of stabilisation of the working point. The 150 Ω resistor is fully decoupled by a 100 μF capacitor, but the 120 Ω resistor is shunted by a comparatively small 4000 pF capacitor. At L.F. and M.F., therefore,

negative feedback is provided by the 120 Ω resistor. This feedback is progressively removed at the H.F. end of the frequency range because the falling reactance of the 4000 pF capacitor causes the 120 Ω resistor to be increasingly decoupled.

(d) An inter-I.F. beat frequency rejector circuit ($L_r C_r$) is included in the cathode to develop a high degree of negative feedback at this unwanted frequency.

(e) A low-frequency (anti-flutter) filter $C_c R_c$, is inserted in the coupling path to the c.r.t.

(f) The screen-grid is decoupled by both a 50 μF and a 5000 pF capacitor. This ensures adequate decoupling over the video frequency range. The parallel 5000 pF capacitor is necessary because the 50 μF electrolytic capacitor, though providing adequate decoupling at low and medium frequencies, offers a significant inductive impedance at the higher frequencies present in the video signal and fails to decouple the screen grid effectively at these frequencies. A capacitor, particularly an electrolytic of large value, is not so simple electrically as its symbol suggests and is in fact a complex impedance with capacitive, resistive and inductive components. Over a wide frequency range it does indeed offer a predominantly capacitive impedance, but at high frequencies this can change to a net inductive impedance, or even, at a certain frequency, it may become purely resistive. This latter resonant effect, where capacitive and inductive components balance out, may occur at more than one frequency, due to the unexpectedly complex nature of the effective equivalent circuit of what appears on a circuit diagram as '*just a simple capacitor*'.

It also often happens that an electrolytic capacitor, which may be part of a two or more section unit, is mounted at a position on the chassis which is remote from the circuit it serves. In a case such as the one discussed here, the lead from the video amplifier screen grid to the electrolytic capacitor may therefore possess significant inductance and, by wiring the second H.F. by-pass capacitor directly between the valve socket and chassis, trouble from this source also is obviated. Such a lead may easily pass near to, say, a line timebase circuit, where high amplitude pulses are all too easily picked up, or it may itself couple video signal information to other parts of the circuit.

Of course electrolytic capacitors are not always shunted by H.F. by-pass capacitors in this way, since it is often not found necessary; perhaps because the component has been so designed that its inductive component is minimised. However, such parallel combinations of large value and small value capacitors do appear from time to time in various parts of radio circuitry and it is as well to appreciate the reason for it.

(g) Suppression of both line and field flyback lines is achieved, by feeding high amplitude negative going pulses to the c.r.t. grid when line and field flyback are taking place. These pulses, which are derived from the appropriate timebases, bias off the c.r.t. beam-current during the flyback periods, so preventing the appearance on the screen of return-path traces.

Fig. 7.2 shows a circuit used in a British 405/625 dual-standards receiver. Note the following points:

(a) The anode circuit includes both a shunt peaking coil ($L_{px}$) and a series peaking coil ($L_{py}$) to give the required H.F. compensation. The series coil is shunted by a damping resistor to reduce its $Q$-factor.

(b) The cathode circuit inter-I.F. beat frequency rejector is switched to tune to 3·5 Mc/s on 405 lines and 6 Mc/s on 625 lines. This is brought about by changing the tuning capacitance from 300 pF on 625 lines to 860 pF on 405 lines.

(c) On 405 lines cathode bias is developed by cathode current, plus a bleed current, through the two 100 Ω resistors in series. One of these resistors is fully decoupled, and the other, which is shunted by a 2600 pF capacitor, is used to give cathode H.F. compensation. On 625 lines the bias is reduced by switching in a 68 Ω resistor to shunt the total of 200 Ω used on 405. The reasons for such a reduction in bias were fully discussed in Chapter 5 on Vision Detectors. The 68 Ω resistor is decoupled by an additional 4000 pF capacitor in parallel with the 2600 pF already in circuit for 405-line working. This again indicates a measure of H.F. compensation.

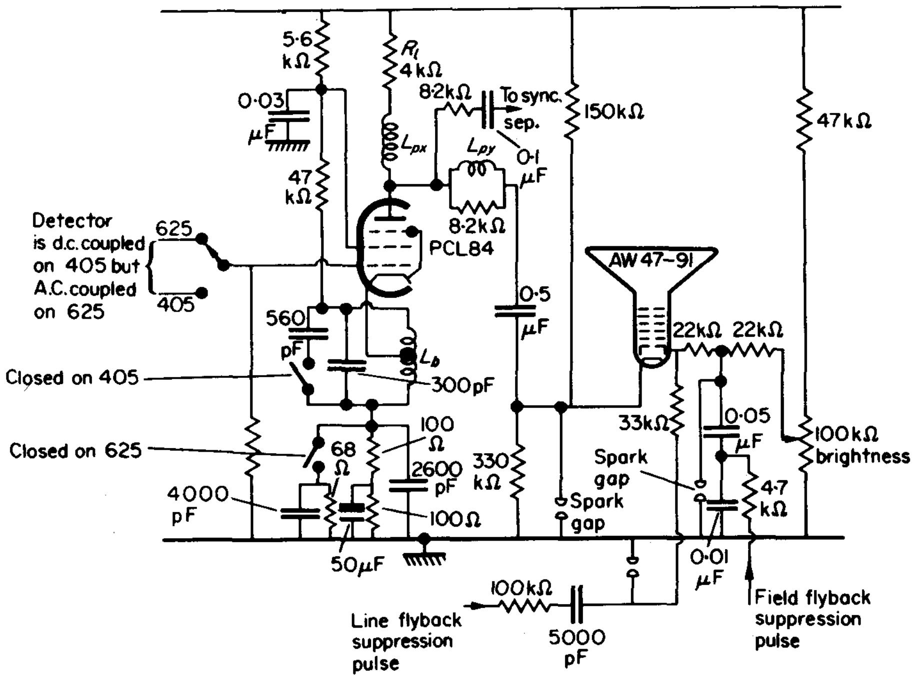



Fig. 7.2 Video amplifier used in a British dual-standards receiver

(d) A.C. coupling is used between the video amplifier and the c.r.t. In this case therefore, the d.c. component is lost. An advantage of this system of coupling in a dual-standards receiver is that the c.r.t. working conditions are unaffected by the change of bias to the video amplifier. The mean c.r.t. cathode potential is no longer determined by the steady voltage at the anode of the video amplifier. The cathode is returned to the junction of two resistors which form a potential divider across the H. T. line. Under normal working conditions the c.r.t. grid voltage will be set by the brightness control so that the grid is negative with respect to the cathode, to place the working bias at a suitable point beneath the Ib/Vg characteristic. The video signal then excurses on either side of this bias potential. This differs from the situation when d.c. coupling is used. In that case the c.r.t. is biased back nearer to the foot of the characteristic, and the video signal always excurses on *one* side only of this bias line; i.e. the sync. pulses 'sit upon' the bias line. As the video signal moves away from the sync. pulse tip level, in the direction of peak white, the net

grid-cathode voltage becomes progressively less negative. Other repercussions of a.c. coupling as opposed to d.c. coupling, were considered in Chapter 5.

(e) Three spark gaps are shown in the grid and cathode circuitry of the c.r.t. Two further spark gaps (not shown on the diagram), are used in this particular circuit. These latter two are connected between the first anode and chassis, and the focus electrode and chassis. The spark gaps are one method of preventing 'flash-over' within the tube from causing damage to external receiver circuitry. The phenomenon of 'flash-over' is unpredictable and its causes are rather obscure. It appears to be due to the build up of very high voltages between the final anode and a neighbouring electrode; e.g. the focus anode. These voltages arise from the accumulation of charges on the electrodes and when 'flash-over' occurs there is a sudden discharge current between the electrodes. During the very brief discharge period, which lasts for perhaps 0·1 µs, high potentials, reaching peak values of several thousand volts, can appear at the c.r.t. base terminals. The remaining discharge path to chassis will be routed through sections of the receiver circuitry (e.g. the video amplifier and detector circuits), unless some easier route, such as the spark gaps of this circuit, is available.

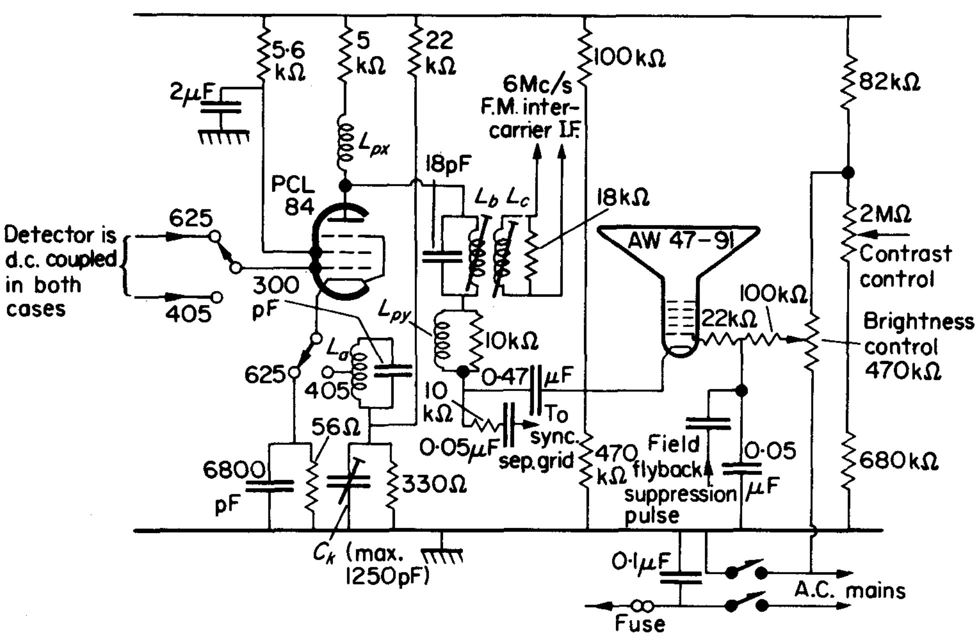



Fig. 7.3 Example of a video amplifier (dual 405/625 standards) using a.c. coupling to the c.r.t.

A rather similar circuit is shown in Fig. 7.3. Note that:

(a) Combined shunt-series H.F. compensation is again used in the anode circuit, with $L_{px}$ and $L_{py}$ as the shunt and series peaking coils respectively.

(b) Cathode H.F. compensation is also used but in this case the entire bias resistance is employed for negative feedback, instead of only a part, as in the previous two circuits. On 405 the compensation capacitor $C_k$ is adjustable. The normal technique is to increase

this until indications of ringing appear on the edges of picture detail, and then to back it off slightly. The cathode bias resistor on 405 is of 330 Ω. This resistor also forms part of a potential divider across the H.T. line so that, as in the previous two circuits, the bias is less dependent upon the valve cathode current. As noted on the diagram, d.c. coupling is used from the detector for both signals and the cathode bias is accordingly reduced on 625 by switching in a 56 Ω resistor in place of the 330 Ω. Clearly a larger capacitance is needed to decouple the 56 Ω resistor, and on 625 the cathode capacitor is increased to 6800 pF. This value again indicates the existence of H.F. compensation since it clearly does not decouple the 56 Ω resistor at L.F. and M.F., but only removes the feedback at H.F.

(c) Included in the series coupling path between the video amplifier anode and the c.r.t. cathode is the primary ($L_b$) of a tuned 6 Mc/s I.F. transformer. The purpose of this transformer is to extract the F.M. inter-carrier sound I.F. of the 625 signal. A bandwidth of some 200 kc/s is necessary for the F.M. signal, and the 18 kΩ resistor across the secondary coil $L_c$ is there to damp the circuit in order to give this bandwidth. The primary tuned circuit also serves another purpose. Since it is, in fact, a parallel tuned circuit connected in series with the signal path to the c.r.t. cathode, it presents a high impedance at 6 Mc/s and prevents the inter-carrier I.F. from reaching the tube. In this way it achieves the same purpose as the 6 Mc/s rejector circuit in the cathode of the circuit of Fig. 7.2. It should be mentioned that in the latter circuit the F.M. sound I.F. is taken off from the detector stage.

(d) A.C. coupling to the c.r.t. cathode is again employed.

(e) The brightness control potentiometer is returned to the mains side of the on-off switch, instead of direct to the chassis. The object is to prevent the appearance of a 'switch-off' spot on the screen centre. When a receiver is switched off the timebase circuits stop immediately and the c.r.t. spot assumes a central position. The cathode and grid potentials rapidly become equal to chassis potential, since the H.T. voltage disappears. However, the tube cathode remains hot for some while and keeps emitting electrons. At the same time, the E.H.T. capacitance formed by the aquadag coating of the tube does not immediately discharge since the resistance of the E.H.T. circuit is very high. For a few moments, therefore, beam current continues to flow, with zero grid-cathode bias and no deflecting fields. A bright spot appears on the screen centre, which can in due course burn the centre phosphor and which in any event often causes concern to set users. Various methods are available to prevent this effect. The method in this circuit is simple and involves no extra components. When the receiver is switched off (at the set), the c.r.t. grid is left connected to the mains neutral line. At the instant of switching off, therefore, the lower end of the brightness control is disconnected from the chassis. Momentarily this takes the c.r.t. grid sharply up to H.T. potential, and a heavy beam current passes for a brief period of time, to discharge the E.H.T. capacitance. This happens *as* the normal H.T. voltage is decaying, and *before* the raster finally collapses, so that the heavy beam current is spread out over the screen face and not concentrated into a central spot. Under some circumstances it is also possible for there to be a 50 c/s voltage applied to the c.r.t. grid at the moment of switching off, and this, if it occurs, further accelerates the discharge of the E.H.T. capacitance. A series grid resistor prevents excessive grid current. A disadvantage of this method is that it conflicts with the British Standards Institution's recommendation that a receiver on-off switch should break both the live and the neutral connections to the mains.

Fig. 7.4 shows a further example of a dual-standards circuit. The diagram looks rather complicated and difficult to follow. It contains some interesting features, however, and to make it easier to look at these, the circuit is re-drawn in Fig. 7.4(b) and Fig. 7.4(c), showing the effective circuits on 405 and 625, respectively. On 405 Fig. 7.4(a) and Fig. 7.4(b) show that:

(a) Shunt H.F. compensation is employed in the anode circuit, by means of $L_{p2}$.
(b) A low-frequency filter ($C_c$, $R_{c_1}$, $R_{c_2}$) and partial d.c. coupling is used between the video amplifier anode and the c.r.t. cathode.
(c) A 3·5 Mc/s inter-I.F. beat frequency rejector is present in the cathode. This is formed by $L_a$, $L_b$, $C_a$ and $C_b$.

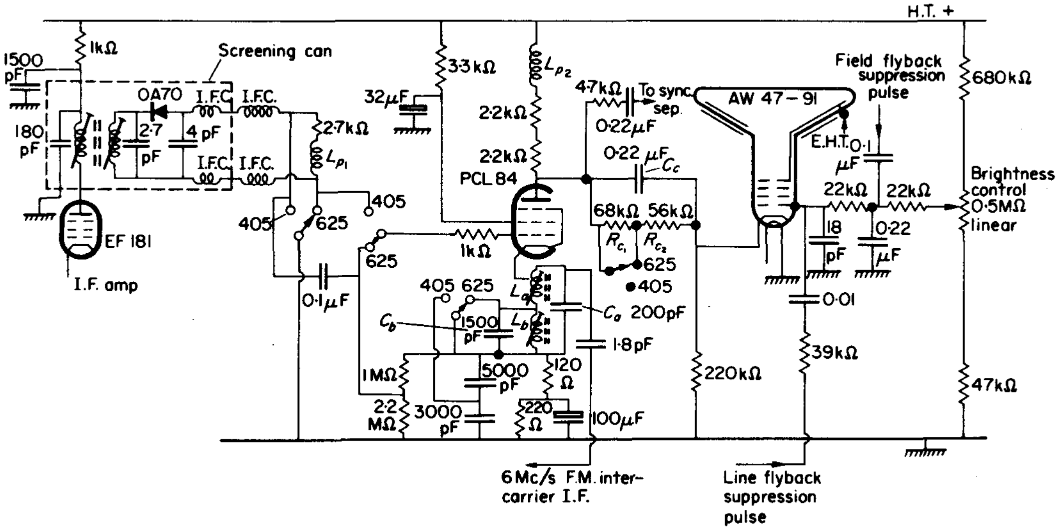



Fig. 7.4(a) Detector and video amplifier circuit in a dual-standards 405/625-line receiver

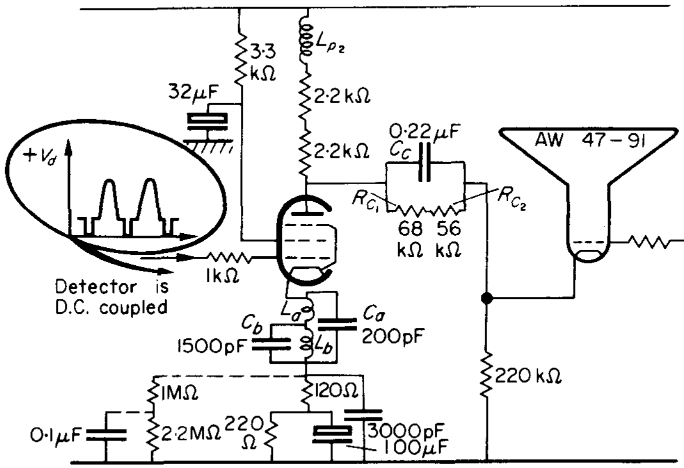



Fig. 7.4(b) Circuit of Fig. 7.4(a) when it is switched to receive the positively-modulated 405-line signal

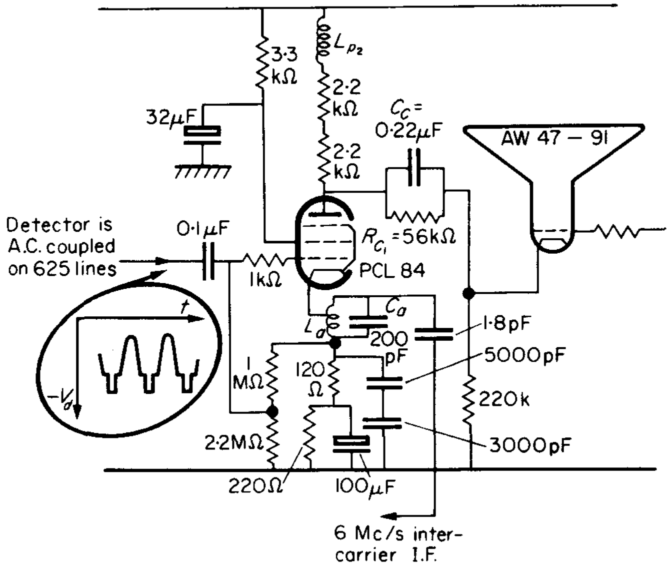



Fig. 7.4(c) Circuit of Fig. 7.4(a) when it is switched to receive the negatively-modulated 625-line signal

(d) The total bias resistance is of 340 Ω. Of this total, one resistor of 220 Ω is fully decoupled, and H.F. compensation is given by the remaining 120 Ω resistor, which is only partly decoupled at L.F. and M.F.

(e) The dectector is d.c. coupled.

On 625, Fig. 7.4(a) and Fig. 7.4(c) show that the following changes are made:

(f) The detector is a.c. coupled.

(g) The video amplifier grid returns to the junction of two resistors which form a potential divider across the cathode bias voltage. The resistor values show that approximately one-third only of the available cathode voltage is used to bias the valve on 625. The working point thus moves forward from its position near cut-off on 405 to a point beneath the centre of the $I_a/V_g$ characteristic on 625. This is logical, since the detector is d.c. coupled on 405 but a.c. coupled on 625.

(h) The cathode tuned circuit is modified by short circuiting $L_b$ and $C_b$, to leave $L_a$ tuned by $C_a$. This forms a 6 Mc/s circuit across which the F.M. inter-carrier sound I.F. is developed. The 6 Mc/s signal is taken-off via a 1·8 pF capacitor to the 6 Mc/s sound I.F. amplifier. At the same time, the presence of the tuned circuit in the cathode fulfils its normal purpose of preventing the inter-I.F. signal from going forward to produce a dot pattern on the screen.

(i) The reduction of bias on the video amplifier valve on 625 causes a drop in the 'no-signal' voltage at the anode. Since partial d.c. coupling is used, the c.r.t. cathode must follow this voltage change. To compensate for this one of the two resistors in the low-frequency filter is switched out. The potential divider between anode and chassis is now made up of a 56 kΩ and a 220 kΩ resistor, instead of a 124 kΩ (68 kΩ + 56 kΩ) and the 220 kΩ resistor. The effect of this change in the potential divider values is to lift the c.r.t. cathode potential back again by the appropriate amount. As a result of this, the c.r.t. cathode rests at approximately the same potential on 625, as it does on 405.

Fig. 7.5 shows another 405/625 dual-standards receiver circuit. Note the following:

(a) A.C. coupling from the detector to the video amplifier is employed on both systems. As explained in Chapter 5, this removes the need to change the bias on the video amplifier when switching from one system to another.

(b) The video amplifier valve is an R.F. pentode feeding the c.r.t. via a cathode follower. The anode load resistor of 10 kΩ is higher than in any of the circuits so far looked at. This is possible, as explained in the previous chapter, because the presence of the cathode follower off-loads from the video amplifier anode a substantial proportion of the total circuit shunt capacitance ($C_t$).

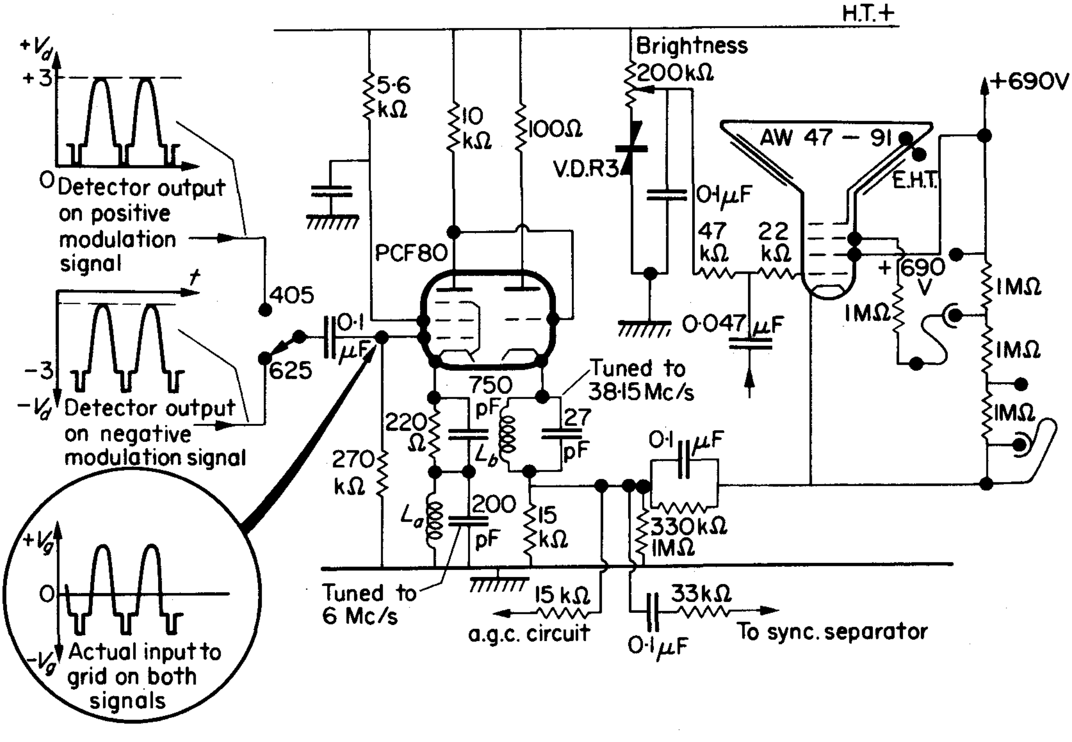



Fig. 7.5 Video amplifier with cathode-follower in a dual-standards receiver

(c) The cathode follower load resistor is of 15 kΩ; a value chosen so that the triode can develop the same peak signal voltage as is present at the pentode anode, but with a smaller anode current than is passed by the latter.

(d) A measure of H.F. compensation is evident in the pentode cathode. The 220 Ω bias resistor is decoupled by a capacitor of only 750 pF which indicates that negative feedback must be developed at L.F. and M.F. and a relatively higher gain is achieved at H.F.

(e) A 6 Mc/s inter-carrier beat frequency rejector is included in the pentode cathode. The take-off point (not shown in this diagram), for this 6 Mc/s F.M. sound I.F. signal is in the vision detector circuit. $L_a$, with its associated 220 pF capacitor, removes any remaining component at this frequency which may reach the video amplifier.

(f) Placed in the triode cathode is an I.F. trap circuit tuned to 38·15 Mc/s (i.e. the sound I.F.

when switched to 405). The a.g.c. circuit is fed from this cathode follower stage, and evidently the trap circuit was found necessary to reduce the possibility of sound-on-vision.

(g) The 100 Ω resistor in the anode of the cathode follower stage is included to prevent possible instability due to back-coupling via the H.T. line. Many similar circuits do not include this resistor.

(h) Voltages on the c.r.t. cathode and grid are worth attention. The brightness control allows a range of adjustment of grid potential from approximately 60 to 210 volts, positive to chassis. The c.r.t. cathode rests at +190 V. This results mainly from the C.F. current through the 15 kΩ resistor and the beam plus the boost H.T. line bleed currents through the 330 kΩ resistor. A step adjustment is available by means of a fly-lead, so that in the event of the c.r.t. being replaced, a satisfactory range of brightness may be achieved. This adjustment is made available because the cut-off point of a replacement c.r.t. may be different from that of the original.

(i) Suppression of the switch-off spot is brought about in this circuit by means of a voltage-dependent-resistor (V.D.R.3) which forms the lower arm of the brightness control potential divider. When the receiver is switched-off and the H.T. voltage disappears, the resistance of the v.d.r. immediately becomes very high. This allows the charge on the associated capacitor to remain for a short time so that the grid is momentarily positive with respect to the cathode. The result of this is that a high beam current passes for a brief instant and this discharges the E.H.T. smoothing capacitance.

The diagram of Fig. 7.6(a) shows a circuit suitable for the C.C.I.R. 625-line system which has a video bandwidth of 5 Mc/s and an inter-carrier I.F. of 5·5 Mc/s (cf. 5·5 Mc/s and 6 Mc/s for the corresponding standards with the British 625-line system).

The following points should be noted:

(a) The 5·5 Mc/s inter-carrier F.M. sound I.F. is taken off by means of a parallel tuned circuit placed in series with the signal path from the detector to the video amplifier grid. The

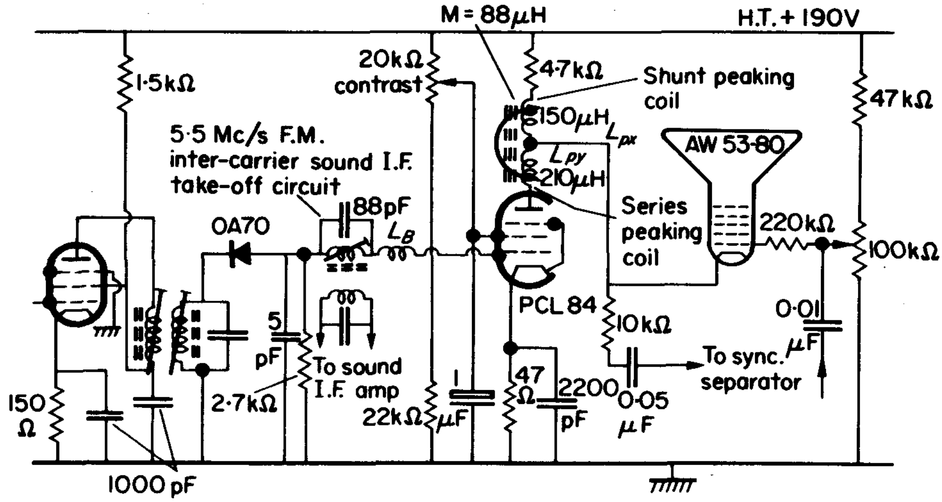



Fig. 7.6(a) Video amplifier for 625-line C.C.I.R. standards compensated by mutually-coupled series and shunt anode coils and by associated differential cathode feedback

tuned circuit forms the primary of a tuned I.F. transformer and serves the additional function of preventing the 5·5 Mc/s beat frequency from being passed through the video amplifier.

(b) The anode circuit contains both shunt ($L_{px}$) and series ($L_{py}$) peaking coils to give H.F. compensation. Unlike the circuits so far studied, these two coils are in fact mutually coupled and give a much enhanced H.F. response.

(c) Cathode H.F. compensation is also employed, by means of the 47 Ω resistor and 2200 pF shunt capacitor.

(d) For the added instruction afforded, the circuit is redrawn in Fig. 7.6(b) without any compensation. Curve A on Fig. 7.6(d) shows the response curve of the uncompensated amplifier; curve B shows the effect of introducing the cathode compensation, and finally

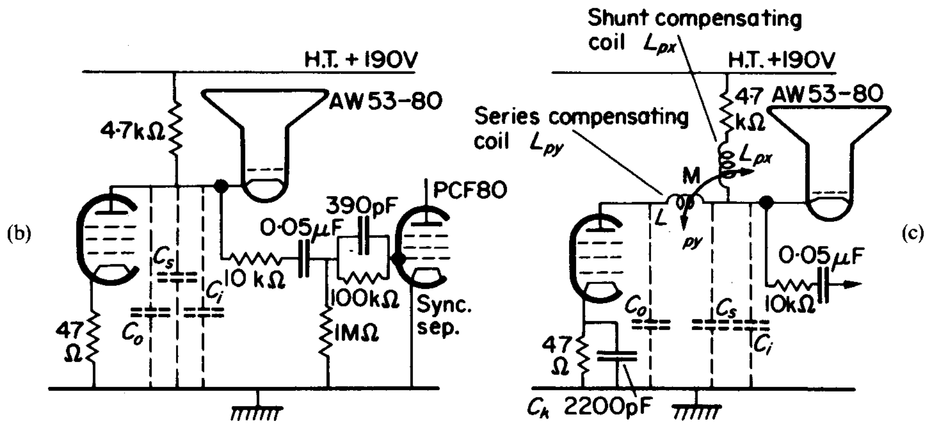



Fig. 7.6(b) Showing the basic uncompensated version of Fig. 7.6(a)

Fig. 7.6(c) Showing how the series coil $L_{py}$ 'stands-off' the capacitance ($C_s+C_i$) pF and forms a low-pass filter terminated by $L_{px}$ in series with the 4·7 kΩ load resistor

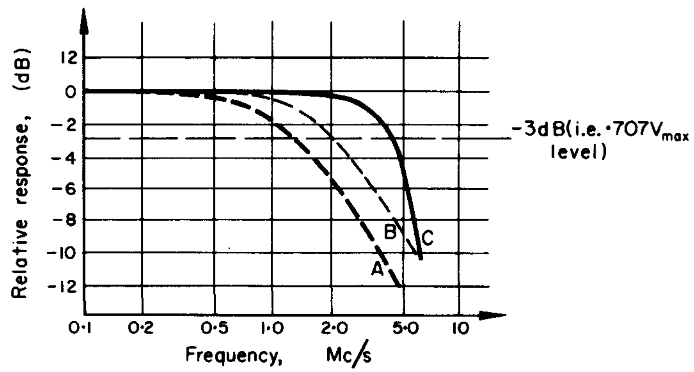



Fig. 7.6(d) Response curves for the video amplifier shown in Fig. 7.6(a)

A Uncompensated as in Fig. 7.6(b).
B Effect of introducing differential feedback by shunting the cathode resistor by $C_k = 2200$ pF.
C Combined effect of anode and cathode compensation.

curve C indicates the performance when both the cathode and anode compensation circuits are included. It is interesting to note that the $-3$ dB frequency ($f_2$) moves up from below 1·5 Mc/s without any compensation, to just under 5 Mc/s with the full compensation. This shows that a very substantial improvement in performance is gained by the inclusion of these techniques.

The circuit of Fig. 7.7 shows a video amplifier for a 625-line negative-modulation receiver. Note that:

(a) The video amplifier is one section of the PFL200 double-pentode valve. This is a power pentode which is able to set up an adequate signal voltage across a comparatively low value anode load resistor. It has a frame grid, which gives it a high mutual conductance of some 20 mA/V at 30 mA. The other section of the valve is a medium slope voltage amplifying pentode suitable for use in various other parts of the receiver, e.g. as a sync. separator.

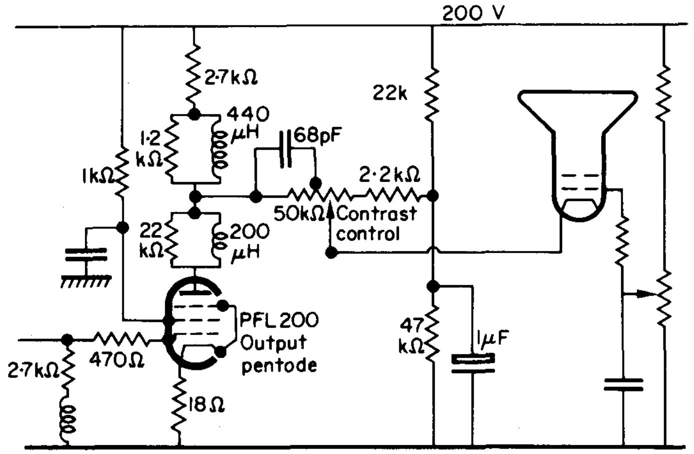



Fig. 7.7 Video amplifier using a 'high-level' contrast control (negative-modulation receiver)

(b) A small undecoupled resistor in the cathode circuit sets the working point back a little way from $V_g = 0$, and so prevents grid current on peak-white. It also gives a small measure of negative feedback.

(c) Both series and shunt H.F. compensation peaking coils are employed. These are both damped to reduce their $Q$-factors to an acceptable level.

(d) An interesting feature of the circuit is the 'high-level' contrast control. Contrast control in many receivers is achieved by variation of the net a.g.c. bias voltage to the I.F. amplifier stages. In this receiver the contrast control takes on something of the appearance of an A.F. gain control in a normal sound receiver. The control is seen to form a potential divider across the anode load circuit of the video amplifier. As the slider moves to the left (i.e. as the contrast control is turned up), an increasing proportion of the available video signal voltage is fed to the c.r.t. cathode. It is evident that changing the position of the control alters the d.c. level of the c.r.t. cathode. However, the right-hand end of the contrast control is returned to the junction of a 22 kΩ and a 47 kΩ resistor. These resistors form a potential divider across the H.T. line and the values are so chosen that the junction

voltage is equal to that which exists at the video amplifier anode during black level. This means that during any time when the video signal is at black level, there is zero potential between the ends of the contrast control. Thus, at every setting of the contrast control, the d.c. level of the slider, and hence of the c.r.t. cathode, is constant when the signal is at black level. The c.r.t. beam current is therefore modulated in the normal way by the excursions of the video signal voltage, from black level towards peak white level. For a given incoming signal, the effect of turning up the contrast control is simply to increase the amplitude of the video signal applied to the c.r.t., without causing any variation of the beam current corresponding to black level. This method of control is referred to as a 'high-level' contrast control because the input to the c.r.t. is controlled by producing a video signal which has a constant maximum amplitude which is greater than that required and then introducing attenuation in the signal path. With the alternative method of control, the amplitude of the modulated vision signal is controlled at a much earlier (lower-signal-level) point in the receiver by varying the gain of the I.F. and R.F. amplifiers. Both methods have their advantages and disadvantages. These will become apparent when a.g.c. systems are studied.

The contrast control potentiometer itself is worth attention. A 68 pF capacitor is shown connected between one end of it and a fixed tap positioned some distance along it. This is to compensate for the loss of some measure of H.F. response due to the presence of part of the resistance of the contrast control in series with the video signal path. The position of the tapping point may vary from one circuit to another. A commonly used potentiometer has the tap at a point 66% along the resistor, but others tapped at 33% or 50% are available.

The circuit of Fig. 7.8 provides a further example of the use of a high level contrast control. Note the following points:

(a) The video signal is direct coupled on the 405-line signal but is only partially d.c. coupled on the negatively modulated 625-line signal. The cathode bias is reduced when switched to the latter signal, but not to the extent which would be necessary if pure d.c. coupling were used.

(b) On 405, a 3·5 Mc/s inter-I.F. beat frequency rejector circuit is present in the cathode.

(c) A measure of H.F. compensation is evident in the cathode since the resistors are not fully decoupled.

(d) On 625 the 6 Mc/s inter-carrier F.M. sound I.F. signal is taken out by means of a parallel tuned circuit placed in series with the anode load circuit. The signal is passed to the 6 Mc/s I.F. amplifier via a 3·3 pF capacitor.

(e) The second anode inductor $L_p$ acts as a series peaking coil. In this circuit it is damped by a comparatively low value 4·7 kΩ resistor to prevent any tendency for it to resonate with its own shunt capacitance.

(f) A 35 kΩ potentiometer, connected in parallel with the 4·7 kΩ anode load resistor, provides a simple method of achieving contrast control. As before, the small (22 pF) capacitor in shunt with a section of the contrast control, reduces the extent to which the H.F. response is changed as the position of the contrast control slider is moved. The section of the potentiometer between the slider and the junction with the anode load resistor lies in series with the video signal path to the c.r.t. cathode. This series resistor acts in conjunction with the total stray capacitance between the slider and chassis, as a low-pass type filter. (Compare the I.F. filter in a standard detector circuit.) At the H.F. end of the video signal spectrum this filter would result in noticeable attenuation, and to prevent

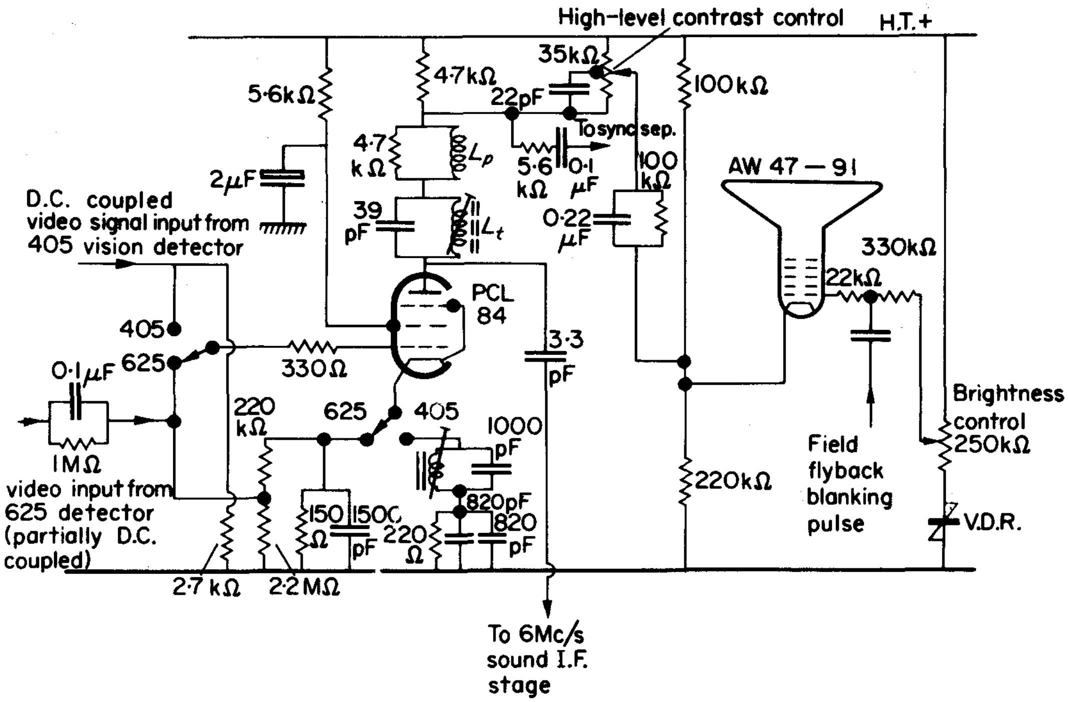



Fig. 7.8 British dual-standards video amplifier employing a 'high-level' contrast control

this the series impedance is brought down at H.F. by the small shunt capacitor. The tapping point is fixed to give the best overall response.

(g) A low frequency anti-flutter filter, consisting of a 0·22 μF. capacitor and 100 kΩ resistor, is included in series with the signal path to the c.r.t. cathode.

(h) A voltage-dependent-resistor (v.d.r.) is included in series with the brightness control to maintain the brightness control voltage more nearly constant and also to assist in removing the switch-off spot in the manner previously described. (See (i) on page 125.)

(i) A negative-going field flyback suppression pulse is fed to the c.r.t. grid to take out the field flyback lines. This is derived from the output stage of the field timebase.

# 8

# Video Amplifiers: Transistor Circuits

This chapter deals with the application to transistor circuits of the principles discussed in the previous two chapters. Basically a transistor video amplifier performs the same function as its valve counterpart, and the general form of the circuitry is recognisably similar in the two cases. There are of course differences made necessary by the nature of the transistor, and attention will be drawn to these. It is essential for a technician to be just as adept at thinking in terms of transistorised circuitry as of valve circuitry. The greatest aid to understanding what is going on in transistor circuits is a clear mental picture of the relationship between input and output currents and voltages. Both in this chapter and in the one which follows on synchronisation separators an attempt is made to give perspective to this subject by the inclusion of a number of illustrative diagrams showing the nature of input and output signals. These should be studied carefully.

### Basic circuit arrangements

Three basic circuits are shown in Fig. 8.1. The first point to be noticed about each of them is the provision to the collector of an auxiliary H.T. supply voltage.

Most transistor circuitry functions with a collector supply voltage $(-V_{cc})$* of less than 20 V. The maximum signal *voltage* output available at the collector of a resistance-loaded transistor amplifier obviously cannot exceed the supply voltage. Thus when the collector current $I_c=0$ mA, the collector voltage equals the supply voltage, and when the collector current is such that the volts drop across the load resistor equals the supply voltage (i.e. $I_cR_l=V_{cc}$), then the collector voltage is zero.

Since the peak-to-peak video signal drive voltage for a c.r.t. must be of the order 40 to 90 volts (depending upon the c.r.t.), it follows that the video amplifier stage responsible for feeding the video signal voltage to the c.r.t. must have a supply voltage of a comparable order (e.g. $-50$ to $-120$ V). This calls for a special auxiliary d.c. supply, which is usually provided by rectifying and smoothing the voltage developed across a separate secondary winding provided on the line-output transformer for this purpose. The bulk of the remaining transistor circuits are fed with the normal low voltage supply, either battery or mains derived; so that the $-V_{cc}$ line is typically between $-10$ V and $-20$ V.

The first of the outline circuits in Fig. 8.1(a) shows a single stage common emitter video amplifier. Though such simple single stages are sometimes used, there are disadvantages to them. These are:

(a) The input impedance to a common emitter amplifier is low and loads the detector circuit. This is particularly the case at high frequencies because of feedback from the output to the input circuits through the transistor capacitance (i.e. due to the 'Miller effect').

* British Standard 3363:1961 recommends the use of the symbol $V_{cc}$ for the collector *supply* voltage. The symbol $-V_{cc}$ for the supply line voltage in Fig. 8.1 shows that p.n.p. transistors are in use.

(b) The circuit does not provide the best means of feeding the sync. separator. A transistor sync. separator requires a current input and is better driven from a low impedance source. In valve video amplifier circuits the sync. separator, like the c.r.t., requires a voltage input and is therefore usually fed from the anode of the video amplifier. This is possible without harming the video frequency response of the stage. To feed both the sync. separator and

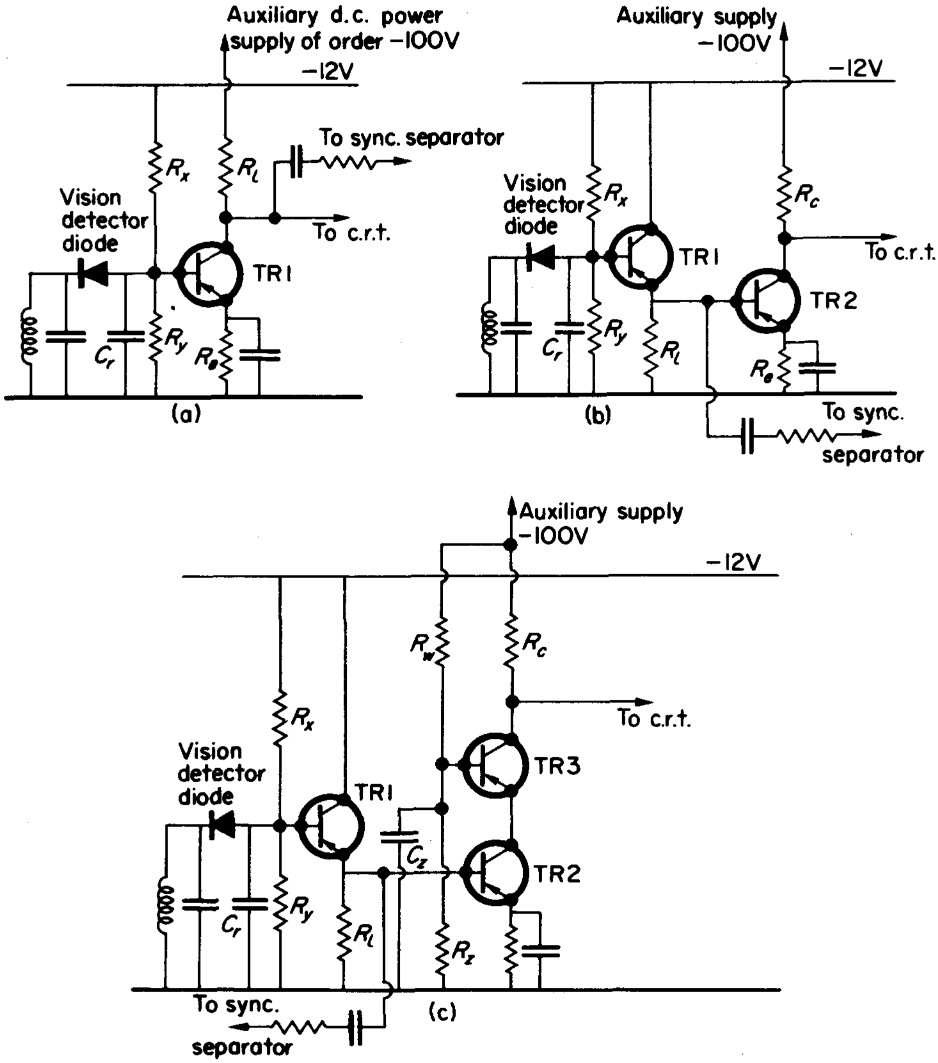



Fig. 8.1 Showing in outline form the three most common video amplifier circuit arrangements found in domestic transistor television receivers

(a) Single-stage common emitter video amplifier.

(b) Two-transistor video amplifier using a common collector (i.e. emitter follower) stage, feeding a common emitter stage.

(c) Three-transistor video amplifier. The emitter-follower TR1 feeds a 'cascode' type amplifier formed by TR2 and TR3.

the c.r.t. from the collector of a common emitter transistor video amplifier presents greater difficulty, although it is none-the-less sometimes done.

The circuit of Fig. 8.1(b) is perhaps the most popular. Here the vision detector feeds an emitter follower (i.e. common collector) circuit which forms the first of a two-stage video amplifier. Such an arrangement has the following advantages:

(1) The input impedance of an emitter-follower is much higher than that of a common emitter stage. (In the same way a valve cathode follower has a higher input impedance than the normal 'grounded cathode' amplifier.) Hence the vision detector is not shunted by a low impedance, and, moreover, there is no difficulty due to feedback through the transistor at high frequencies.
(2) The low output impedance at the emitter of the emitter-follower provides a useful point from which to feed the sync. separator stage.
(3) The thermal stability of the circuit is good.
(4) The low output impedance is readily matched into the low input impedance of the second common emitter amplifying stage.

It should be noted that, like the cathode-follower, an emitter-follower has a voltage gain of less than unity, and also introduces no phase inversion. Essentially it functions in the dual stage video amplifier as a matching device between the vision detector and the common emitter stage. It is the latter which gives the necessary gain, whilst the emitter follower makes it easier to obtain the required frequency bandwidth, whilst at the same time providing a good direct match to the sync. separator.

The third circuit (Fig. 8.1(c)) is similar but uses a two-transistor 'cascode' stage instead of the common emitter transistor. The cascode circuit includes a common emitter amplifier, the collector load of which is the input circuit to a common base amplifier. The circuit compares in form with the cascode valve circuit, often found in tuner-units, where a grounded cathode first valve feeds into the cathode of a second grounded-grid valve.

As a video amplifier, the chief merit of the cascode arrangement is that it allows transistors to be used which, individually, would be incapable of withstanding the collector voltage swing (e.g. of the order 40 to 90 volts peak to peak) necessary to drive the receiver c.r.t. The development of transistors of higher voltage rating, and perhaps of a c.r.t. requiring a smaller signal drive voltage, may make the cascode *voltage sharing* form of circuit unnecessary.

Before going on to study actual examples of commercial transistor video amplifier circuits, some further general aspects of transistor behaviour and circuitry will be considered.

## Comparison of voltage polarities in valve and transistor circuits

In Figs 8.2 and 8.3 video signal waveforms and voltage polarities relative to the chassis are examined for valve and transistor circuits using both grid and cathode modulation of the c.r.t.

Little has yet been said on the subject of methods of modulation of the c.r.t. beam current. The matter is dealt with in more detail later in this chapter. Suffice it to say at the moment that the beam current of a c.r.t. may be increased either by driving the grid less negative with respect to the cathode or by driving the cathode less positive with respect to the grid. The c.r.t. rests with a negative bias between grid and cathode, and the object is to decrease this bias as the video signal moves from black level towards peak white. Obviously, a positive-going video signal to the grid achieves this, or alternatively, a negative-going video signal applied to the cathode has the same effect.

**Grid modulation of the c.r.t.**

The two diagrams of Fig. 8.2 compare valve and p.n.p. transistor video amplifiers connected to grid-modulate the c.r.t. Since a positive-going video signal is needed at the anode (or collector) of the video amplifier, a negative-going video signal is needed at the grid (or base). Reference back to the diagrams of Fig. 2.6 in Chapter 2 shows how this polarity of video signal may be obtained from either a positively or negatively modulated carrier. In the insets on Fig. 8.2 the waveforms are sketched as voltages referred to chassis (zero) potential.

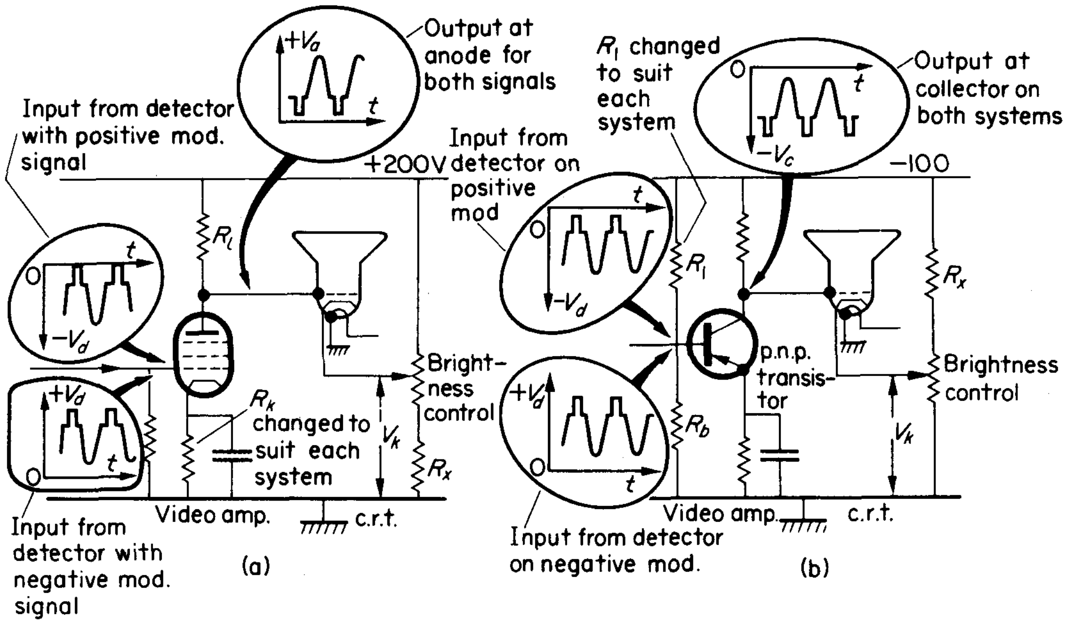



Fig. 8.2 Comparison of video signal voltage polarities in valve and transistor receivers, for both positively and negatively modulated signals, when *grid* modulation of the c.r.t. is employed

Careful attention is now needed to compare conditions in the transistor circuit with those in the valve circuit, firstly for a positively modulated signal, and secondly when a negatively modulated signal is being received.

*Valve circuit on a positively-modulated vision signal* (see Fig. 8.2(a) and Fig. 2.6(a), (d), (e))

Dealing initially with the video signal from the positively modulated carrier, the valve must be biased *forward* near to $V_g = 0$ V so that as the signal moves towards peak-white, the negative voltage on the grid increases and the anode *current falls*. A *fall* in anode current brings about a *rise* in anode voltage relative to chassis. The c.r.t. grid is then carried in the positive-going direction as the picture signal moves from black-level towards peak white.

*Transistor circuit on a positively-modulated vision signal* (see Fig. 8.2(b) and Fig. 2.6(a), (d), (e))

The transistor must be biased *back* towards $I_b = 0$ μA so that as the video signal moves towards peak-white level, the forward bias current increases and the collector current *increases*. As the collector current *increases*, the volts drop across the collector load resistor increases,

and the collector voltage relative to chassis becomes *less negative.* Thus, as the collector current increases, the collector voltage moves towards the chassis potential, which in this case is connected to the positive terminal of the d.c. supply. At the collector, therefore, the required positive-going video signal is available for driving the c.r.t. grid.

Notice that the cathode of the c.r.t. is returned to the brightness control which sets the cathode at a voltage which is less negative with respect to chassis than that at the grid. Brightness control is achieved by varying this net 'positive bias' between cathode and grid. The resistor $R_x$ fixes the minimum limit of this bias and prevents the cathode from being set more negative than the grid (i.e. prevents the possibility of positive voltage on the grid). In the valve circuit the resistor $R_x$ has to be on the opposite side of the brightness control to achieve the same end. The reader should satisfy himself that this is so.

*Valve circuit on negative modulation* (see Fig. 8.2(a) and Fig. 2.6(f), (g), (h))

The video waveform applied to the grid has the same negative-going sense as before, but it exists as an entirely positive voltage relative to chassis instead of one which is set below zero potential. To accommodate this, the bias on the cathode of the video amplifier must be increased so that the sync. pulse tips just fall behind the $V_g = 0$ V point. When the signal moves towards peak-white the net grid-cathode voltage moves more negative and the anode current falls. As before a positive-going output appears at the anode. A change of bias is therefore all that is needed to enable the video amplifier to deal with the changed d.c. level of the video input signal, brought about by switching the detector to deal with the alternative sense of the vision signal modulation.

*Transistor circuit on a negatively-modulated vision signal* (see Fig. 8.2(b) and Fig. 2.6(f), (g), (h))

As with the valve, a change of bias is all that is needed to enable the transistor circuit to handle the negatively-modulated signal. Whereas previously the transistor was biased back near to the $I_b = 0$ μA point, it must now be forward biased so that the sync. pulse tips drive it backwards towards the $I_b = 0$ μA point. Note that the input signal is an 'all-positive' voltage. The least positive level is reached on peak-white, and the transistor passes the greatest current at this level. Once again, maximum collector current means the least negative collector voltage; i.e. the collector signal voltage is positive-going.

**Cathode modulation of the c.r.t.**

In this case a negative-going signal is needed at the anode (and collector) of the video amplifier so that a positive-going signal is needed at the grid (and at the base). In Fig. 8.3 the appropriate valve and transistor circuits are compared.

*Valve circuit on positively-modulated vision signal* (see Fig. 8.3(a) and Fig. 2.6(a), (b), (c))

The valve must be biased back towards the cut-off point so that the positively-going (and 'all-positive' with respect to chassis) video signal drives the anode current to maximum on peak-white. This causes minimum anode voltage on peak-white; i.e. the anode signal is negative-going and hence suitable for feeding to the c.r.t. cathode.

*Transistor circuit on positively-modulated signal* (see Fig. 8.3(b) and Fig. 2.6(a), (b), (c))

Here minimum collector current is required on peak-white so that the collector voltage then has its most negative value. The transistor must be forward biased in such a way that the positive-going voltage applied to the base-emitter junction, causes the net forward bias to

decrease as the video signal moves from black-level towards peak-white level. As the forward bias decreases, so also does the collector current, and the collector voltage becomes more negative with respect to chassis. Notice that brightness control is afforded by variation of the negative potential on the c.r.t. grid. The resistor $R_x$ is now placed at the chassis end of the brightness control to ensure that the voltage on the tube grid does not become *less* negative than the voltage at the cathode. Once again, in the valve circuit, to achieve the same purpose the resistor $R_x$ has to move to the opposite side of the brightness control.

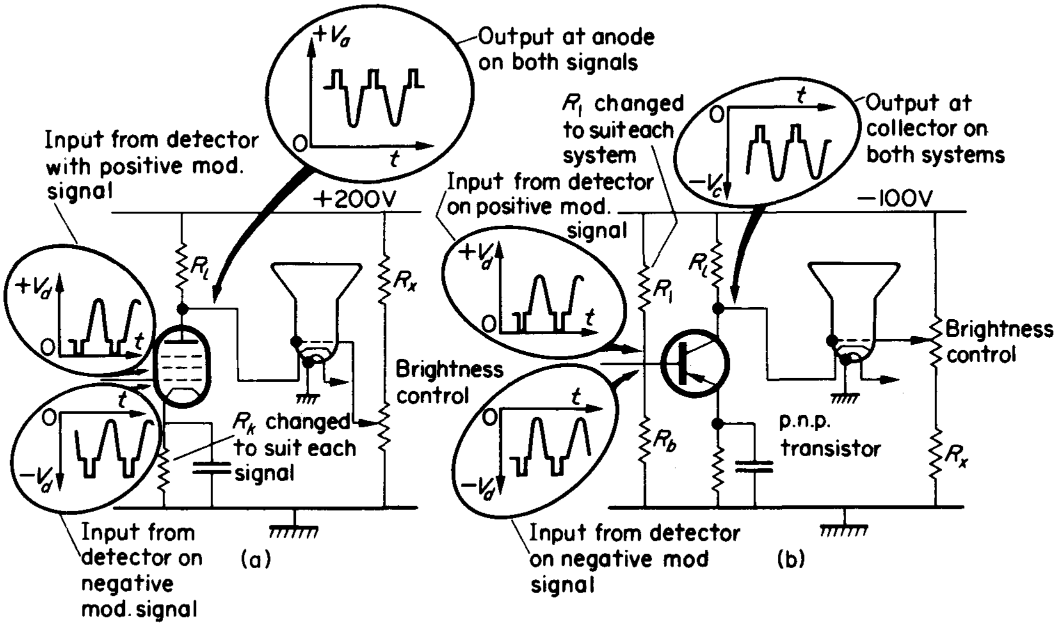



Fig. 8.3 Comparison of video signals voltage polarities, for both positively and negatively modulated signals, when *cathode* modulation of the c.r.t. is employed

*Valve circuit on a negatively-modulated vision signal* (see Fig. 8.3(a) and Fig. 2.6(f), (i), (j))

The valve is here biased up near to the $V_g = 0$ point. On peak-white the net grid-cathode voltage on the valve is only slightly less than this fixed-bias level, moving progressively back further as the signal moves towards the black-level. Hence maximum anode current corresponds with peak-white, and minimum anode current occurs on the sync. pulse tips. The anode voltage waveform is therefore negative going as required.

*Transistor circuit on a negatively-modulated signal* (see Fig. 8.3(b) and Fig. 2.6(f), (i), (j))

The 'all-negative' input waveform reaches its greatest negative level on the sync. pulse tips. The transistor input circuit is therefore biased back towards $I_b = 0$ μA so that on peak-white the base current (and hence the collector current) is only very small, becoming greater as the video signal moves towards black-level. Minimum collector current corresponds to the maximum negative voltage at the collector, so that once again the necessary negative-going collector voltage signal is derived for the c.r.t. cathode.

These results may be summarised as follows:

(1) The sense of the video signal relative to black-level, seen either at the input or the output terminals, is the same in the valve and transistor circuits.
(2) When the valve is called upon to deliver maximum anode current the transistor has to pass minimum collector current, and vice versa.
(3) When the valve works with a minimum negative bias on the grid (i.e. passes greatest current under no-signal conditions), the transistor has its least value of forward bias (i.e. passes minimum current under no-signal conditions), and vice versa.

The reversed conditions evident under (2) and (3) are due to the fact that the p.n.p. transistor works from a negative supply voltage rail, whereas the valve has a positive supply rail. A positive-going movement of potential at the transistor collector has to stem from an *increase* in collector current, whereas a positive-going excursion in the anode potential of the valve stems from a *reduction* in anode current.

It must be stressed that the foregoing argument has been based upon a comparison of a valve amplifier with a *p.n.p.* transistor amplifier. In the case of an *n.p.n.* transistor, a positive supply voltage rail is needed, and a positive drive to the base is necessary to increase the collector current. Obviously, therefore, the working conditions of an n.p.n. transistor video amplifier *are exactly parallel* to those of a valve and no mental gymnastics are needed to compare the two.

The diagrams of Fig. 8.4, with the associated diagrams of Fig. 8.5, are intended to give further experience in studying p.n.p. transistor circuit behaviour. Cathode modulation of the c.r.t. is assumed so that a negative-going video signal must be produced by the complete video

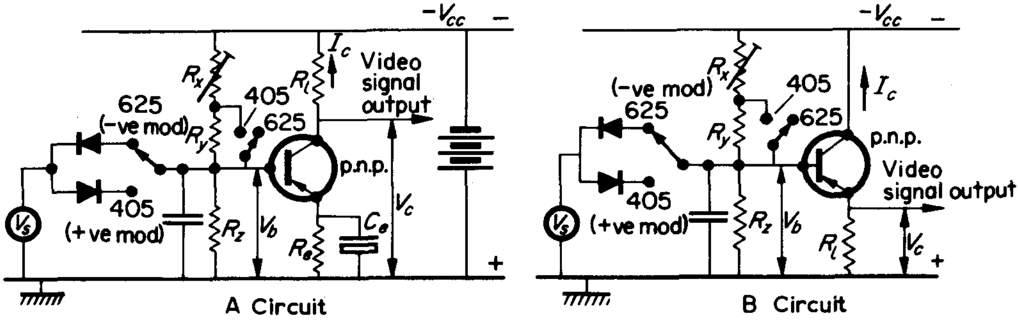



Fig. 8.4 Illustrating basic detector and video amplifier arrangements, on positively and negatively modulated signals, showing at (A) a common emitter circuit and at (B) a common collector circuit. Note that (B) would be followed by a further common emitter stage

amplifier. Fig. 8.4(a) shows an outline circuit of a single common emitter video amplifier of the form introduced in Fig. 8.1(a), whilst Fig. 8.4(b) shows the emitter-follower part of a dual transistor video amplifier of the form introduced in Fig. 8.1(b).

The object in each case is to trace the path of the video signal through from the detector to the output terminals of the transistor, paying particular attention to voltage and current signal waveform polarities on both the positively and negatively modulated signals.

Two columns of waveforms are shown: Figs 8.5(a), (b), (c), (d) and (e) apply to negatively-

modulated signal whilst Figs (a′), (b′), (c′), (d′) and (e′) show conditions when receiving the positively-modulated vision signal.

The first three in each column apply to *both* circuits (A and B) of Fig. 8.4, but (d) and (d′) show the output at the collector of circuit A, whilst (e) and (e′) show the corresponding output at the emitter of circuit B.

The output of circuit B is naturally a positive-going video signal since the emitter 'follows' the base input signal polarity. Reference back to Fig. 8.1(b) shows that this signal then forms the input to the second common emitter transistor of the dual transistor video amplifier. This second stage provides the necessary amplification and inverts the signal to feed the c.r.t. cathode with its required negative-going video signal.

Dealing firstly with the negatively-modulated signal, the successive diagrams in the column must be looked at carefully. The detector delivers a positive-going video signal input to the base of the transistor; see Fig. 8.5(a) and (b). Note that the signal is an 'all-negative' voltage, reaching its greatest negative value on sync. pulse tips.

A p.n.p. transistor base-emitter junction passes an increasing forward current as the base is driven more negative to the emitter. To accommodate the 'all-negative' input signal it is necessary to bias the transistor back towards the cut-off point ($I_b = 0$ μA), so that the base is driven progressively more negative as the video signal voltage moves from peak-white towards the sync. pulse-tip level.

In Fig. 8.5(c) the base current waveform which results from the signal applied from the detector is shown as the input current signal applied to the $I_c/I_b$ transfer characteristic. Minimum collector current is passed during quiescent 'no-signal' times, increasing to a maximum on sync. pulse tips.

In Fig. 8.5(d) the resulting voltage variation at the collector, relative to the chassis potential, is drawn. Once again note how it is that maximum collector current results in the least negative collector voltage. When the video signal is moving from black-level towards peak-white level, the collector current progressively decreases, and the collector voltage moves nearer to the supply voltage of $-V_{cc}$; i.e. a negative-going video signal appears at the collector.

Conversely, at Fig. 8.5(e) the output at the emitter of circuit B shows a positive-going video signal. This follows logically, since maximum emitter current flows on sync. pulse tips and the volts drop across the resistor $R_l$ drives the emitter to its maximum negative level relative to chassis. As the video signal moves from black-level to peak-white, the diminishing emitter current causes the emitter voltage to move in the positive direction towards chassis potential.

It is worth noting that the waveforms at the collector and the emitter are both entirely 'all-negative' with respect to chassis. In terms of the peak-white voltage relative to the black-level voltage, however, the former (collector) delivers a negative-going video signal whilst the emitter delivers a positive-going signal.

The second column of waveforms should now be studied. Here the positively-modulated vision signal gives rise to a detector output which is 'all-positive' relative to chassis; (see (a′)).

Like the waveform on the left at (a), it is a positive-going video signal, but differs from (a) in that it is disposed above chassis potential instead of below it. To accommodate such an input the transistor must be biased well forward (see point X in (b′)) so that as the detector voltage moves more positive (i.e. towards peak-white) the net base-emitter voltage grows less negative and the base current diminishes.

As before, the base current waveform is re-drawn at (c′) beneath the $I_c/I_b$ transfer characteristic. The collector current waveform produced is the same as that at (c); as are the output waveforms at (d′) and (e′).

When switched to the negatively modulated signal.

When switched to the positively modulated signal.

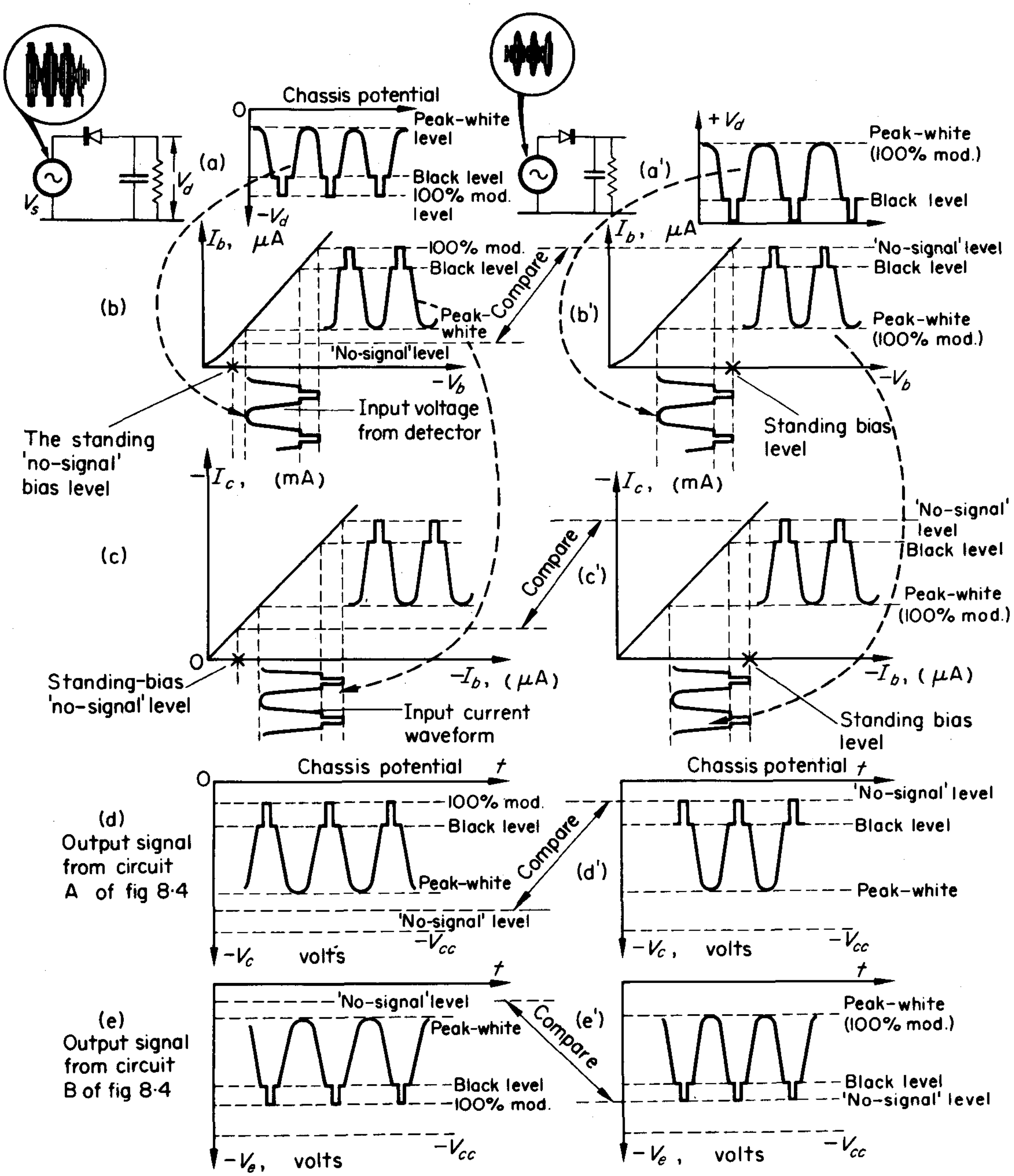



Fig. 8.5 Showing, with particular reference to standing bias levels and to voltage polarities relative to chassis, the conditions in the circuits of Fig. 8.4 when switched to the negatively-modulated 625-line signal (see a, b, c, d, e) and to the positively-modulated 405-line signal (see a', b', c', d', e')

The important difference between the two cases lies in the bias conditions needed, and in the voltages and currents which exist under 'no-signal' conditions. These levels should be compared as indicated on the diagrams.

It has been assumed that cathode modulation is in use in the cases studied. Referring to the diagrams at (d) and (d′) an important point should be noted. With the negatively-modulated signal, the collector rests at its maximum negative level under no-signal conditions; (see (d)). Hence the c.r.t. cathode is also similarly at its most negative level. Thus, under these conditions, the c.r.t. runs at maximum brightness. Naturally this is also true when circuit B is in use, since the voltage waveform at the collector of the second (common emitter) transistor fed by this emitter follower stage must be the same as that shown at (d).

Conversely, diagram (d′) shows that when positive modulation is in use, the video amplifier collector voltage is at its least negative level under no-signal conditions. The c.r.t. cathode is then also at its least negative level and the c.r.t. beam current is a minimum.

The importance of achieving an understanding of signal polarities, and of seeing clearly how currents and voltages vary under the influence of the signal, cannot be stressed too strongly. The student is urged to persist with this subject until satisfied that such an understanding has been arrived at. To assist in thinking through the matter, the following logical steps are listed to provide a basis for reasoning when dealing with p.n.p. transistors:

(1) Determine the polarity of the video signal input which is needed by the c.r.t. (i.e. grid modulation requires a positive-going signal; cathode modulation necessitates a negative-going signal).

(2) Work backwards to the detector to determine what sense of signal is needed from it, in order to arrive at the right polarity at the c.r.t.

(3) Now decide whether the output must be taken from the detector diode's anode or cathode. This will depend upon whether the signal being considered is positively or negatively modulated. The diagrams shown in Fig. 2.6 of Chapter 2 give a clear picture of the possibilities here.

(4) Study the nature of the video signal output produced by the chosen detector circuit. Pay particular attention to the way in which the waveform is disposed relative to chassis potential; i.e. notice if it is 'all positive' or 'all negative' relative to chassis. (N.B. For circuits in which the chassis is connected to the supply negative terminal instead of the positive one, the chassis should *not* be used as the datum line. In this case ignore the chassis and consider voltage variations relative to the supply positive line. This preserves the same picture of voltage variations.)

(5) Having arrived at a picture of the output delivered by the detector, it is now necessary to determine how the transistor has to be biased to allow the video signal to be accommodated under its input characteristic. Clearly, if the signal is entirely positive relative to chassis, then the transistor must be biased well forward (i.e. base negative to emitter) so that when the positive signal is applied, it decreases the net bias. Conversely, if the signal is all-negative with respect to chassis, then the transistor must only have a very small negative forward bias. When the signal is applied the negative voltage between base and emitter then increases with the signal amplitude.

(6) The steady no-signal bias on the transistor is now known. From this the 'no-signal' level of collector current, and then the no-signal level of collector voltage, may be determined. It must be remembered that the collector voltage relative to chassis (or relative to supply positive) is at its most negative level when the collector current is at a minimum, and moves less negative (i.e. in the positive-going direction) as the collector current increases.

(7) Finally, sketches of the collector voltage variation may be made, of the form shown in Fig. 8.5(d) and (e) (or (d′) and (e′)). This is done by showing the disposition of the video signal relative to the no-signal level. Notice that with a video signal derived from a negatively-modulated signal, the level reached on peak-white never reaches the no-signal level because the carrier is never allowed to reach 0% modulation. The output from the video detector is therefore never zero when a signal is present.

It is a useful exercise to think through the same sequence of steps assuming an n.p.n. transistor is used instead of a p.n.p. type, and to decide what modifications need to be made to the arguments of steps (5), (6) and (7) in order to accommodate its opposite polarity voltage and current requirements. Clearly the word 'opposite' gives a clue to the necessary changes.

**Protection against c.r.t. flash-over**

Transistors are easily destroyed by high voltage transient pulses of the kind which appear if flash-over takes place in the c.r.t. Flash-over was discussed earlier when valve video amplifiers were studied. It becomes even more important to take precautionary measures in transistor television receivers.

The most popular method of guarding against these damaging transients is the connection of capacitors between the first and focus anodes, and the earthed aquadag coating around the bulbs of the tube. Such capacitors, which are often of 0·1 μF, must have very short leads and have a high voltage rating. The reason why leads must be short is that the transient waveforms are extremely steep fronted (i.e. they have very fast rise times). The voltage induced across an inductor is proportional to the rate of change of current in the inductor ($e = -L.(di/dt)$). It is sometimes forgotten that even a short straight piece of wire possesses inductance (e.g. 1 in. of 22 S.W.G. wire has an inductance of approximately 0·025 μH).

With the very rapid rate of change of current in external circuits which takes place resulting from flash-over within the c.r.t., large voltages can build up across apparently innocuous leads. Capacitors inserted to provide short circuit paths between the tube anodes and 'earth' will only 'kill' the transient pulse if they are fitted with the shortest possible leads.

**Grid and cathode modulation of the c.r.t.**

Both grid and cathode modulation methods are used in commercial transistor receivers, as indeed they are with valve receivers. The chief points to consider in assessing the relative merits of the two systems are:

(1) The slope of the input characteristic. This is a measure of the beam current change for a given change in video signal drive voltage.
(2) The convenience with which the sync. separator may be fed with its necessary negative-going video signal voltage.
(3) The safety of the c.r.t. in the event of failure of the video amplifier.
(4) The cathode-to-heater voltage stress.
(5) The c.r.t. *no-signal* beam current in a d.c. coupled circuit.

The last point has already been discussed. Minimum *no-signal screen brightness* is achieved on positively and negatively modulated signals using cathode and grid modulation respectively (see pages 73, 74 and 138). The other four matters are now examined.

1. *Input characteristic slope*

The advantage here lies with cathode modulation. The reason is interesting, but not immediately self-evident. When grid modulation is used, the only factor which determines what change in beam current results from a given change in video signal voltage is the change in the grid-to-cathode voltage. No other electrode is concerned with the applied voltage.

When cathode modulation is employed a second factor influences the anode current. This is the voltage between the first anode (i.e. screen-grid) and the cathode. Thus, as the video signal input-voltage moves from black-level towards peak-white, the cathode moves more negative—not only to the control grid but also to the first anode. But the voltage between the first anode and the cathode, *itself* has a marked influence upon the beam current of a c.r.t., so that in addition to the change in beam current brought about by the effective *reduction* in grid-to-cathode *negative* bias, there is a further increase due to the *increase* in the *positive* voltage between the first anode and the cathode.

With grid modulation, the voltage between the first anode and the cathode remains constant. It follows that for a given change in the input signal voltage, the change of beam current is less with grid modulation than it is for cathode modulation. The diagrams of Fig. 8.6 illustrate the point. Fig. 8.6(a) represents the normal $I_b/V_g$ curve for a given c.r.t. It is labelled $V_{a1} = x$ volts, where $x$ is the normal first anode voltage recommended for the tube. Beneath the curve is

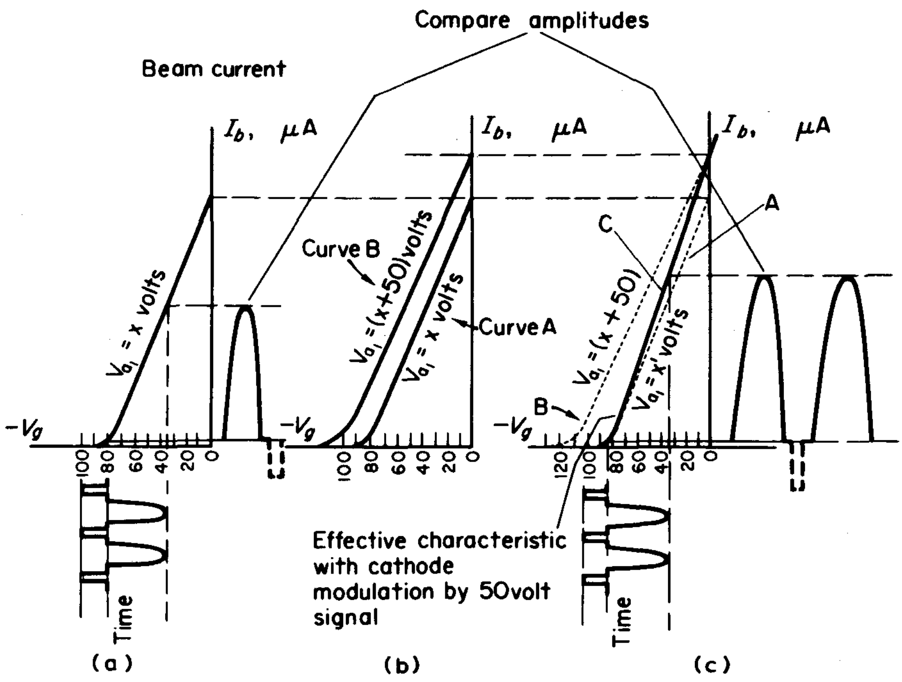



Fig. 8.6 Illustrating how the use of cathode modulation increases the effective slope of the $I_b/V_g$ characteristic of a c.r.t., because the voltage between the first anode and cathode increases with the signal amplitude

(a) $I_b/V_g$ characteristic of a c.r.t. working with grid modulation.

(b) Showing the effect on the $I_b/V_g$ characteristic of an increase in the first anode voltage from $V_{a_1} = (x)$ V to $V_{a_1} = (x + 50)$ V.

(c) Showing the effective $I_b/V_g$ curve when a signal (in this case of 50 V amplitude) is applied to the cathode.

drawn a video signal, the picture voltage of which excurses through 50 volts between black-level and peak-white. The corresponding beam-current waveform is shown. In Fig. 8.6(b) the effect on the $I_b/V_g$ curve of increasing the first anode voltage from $x$ volts to $(x+50)$ V is shown. The new $I_b/V_g$ curve has the *same slope* but moves to the left. These therefore represent two 'static' $I_b/V_g$ curves for two different values of first anode voltage $V_{a1}$.

When cathode modulation is used, however, neither curve on its own represents the true relationship between beam current and applied signal voltage. When the video signal is at black-level, the effective voltage between the first anode and cathode is still $x$ volts, and curve A shows the $I_b/V_g$ relationship. At peak-white, however, assuming the same 50 V signal as before, the cathode is driven a further 50 V negative with respect to the first anode. Hence at this limiting condition the first anode voltage relative to the cathode is now $(x+50)$ V and the effective $I_b/V_g$ curve is given by curve B. It follows that the true dynamic relationship between $I_b$ and $V_g$ is given by a curve whose foot rests at the foot of curve A, and which crosses the $I_b$ axis at the top of curve B.

In Fig. 8.6(c) curves A and B are shown dotted and the effective $I_b/V_g$ curve, shown as curve C, is drawn between them. Having arrived at the actual apparent relationship between applied signal voltage and beam current, the video signal is now shown as a voltage excursion beneath this curve. Note that the curve is still drawn as an $I_b/V_g$ curve and the video signal is shown as a positive-going signal beneath it.

When a signal is applied between the grid and cathode of a c.r.t. (or a valve), it is customary to show it as a variation of grid voltage relative to cathode even when the signal is actually applied to the cathode instead of the grid. This is logical since it is in fact the grid which controls the density of the electron beam, and it would only confuse the issue to plot families of $I_a/V_k$ curves as well as the more widely familiar $I_a/V_g$ curves.

Finally the total change in beam current shown at Fig. 8.6(c) should be compared with the corresponding change in Fig. 8.6(a). The input is the same in each case but the beam current is some 20% greater on peak-white in the cathode modulation case.

2. *Feed to the sync. separator*

As fully described in the next chapter, the most widely used sync. separator circuits employ either a valve or a transistor, which is cut off during the picture information part of the video signal, but is driven sharply into strong conduction by the sync. pulse tips. In the valve sync. separator circuit, the video signal presented to the control grid must therefore have the sync. pulses as the most positive part of the waveform. A negative-going video signal is thus needed by the sync. separator.

If the c.r.t. is cathode modulated, the required negative-going video signal, developed at the anode of the video amplifier for transmission to the c.r.t. cathode, is of the right polarity and order of amplitude which is needed to drive the sync. separator. Hence cathode modulation of the c.r.t. has the added advantage over grid modulation of automatically providing the right polarity of video signal needed to drive the sync. separator. As will be seen from a study of circuit diagrams, the great majority of valve television receivers have the sync. separator fed from the anode of the video amplifier. The only precaution needed is to ensure that the input capacitance of the sync. separator does not shunt the video amplifier anode, and to prevent this a 'stand-off' resistor is used in series with the signal path to the sync. separator grid.

In transistor sync. separators using p.n.p. transistors, however, if the transistor is to be fired into conduction by the sync. pulses and cut off by the picture detail part of the video signal, it follows that the sync. pulses must be the most *negative* part of the signal. Hence, unlike the

valve sync. separator, a positive-going video signal is now needed. If the c.r.t. is *grid* modulated, the necessary positive-going video signal is readily available from the collector of the video amplifier. In the case of a single transistor video amplifier feeding the grid of a c.r.t., the sync. separator may be fed from the collector. If cathode modulation of the c.r.t. is employed, and a dual-transistor video amplifier of the type discussed is used, the necessary positive-going video signal for driving the sync. separator, is readily available at the emitter of the first emitter-follower stage.

The implications of these remarks may be summarised as follows:

(a) With valve receivers, cathode modulation of the c.r.t. is a distinct advantage since the sync. separator is conveniently fed from the video amplifier anode with its required negative-going video signal.
(b) With transistor receivers using a single stage transistor video amplifier, grid modulation of the c.r.t. is more convenient because the positive-going video signal needed for driving the base of a p.n.p. transistor sync. separator is then available at the collector of the video amplifier.
(c) In transistor receivers which employ a dual-transistor video amplifier, the question of feeding the sync. separator is no longer a paramount issue since, whatever the form of c.r.t. modulation, video signals of both negative and positive polarities are always available. Thus if cathode modulation is used, there is a positive-going video signal available at the emitter of the first emitter-follower transistor, but if grid modulation is used there is a positive-going signal at the collector of the second (common-emitter) transistor.

3. *Safety of the c.r.t. in the event of video amplifier failure*

With valve receivers, the advantage again lies with cathode modulation. This is evident from a study of the diagrams Fig. 8.2(a) and Fig. 8.3(a). Should the emission of the video amplifier valve in Fig. 8.2(a) fail, there is no volts-drop across $R_l$, and the tube *grid* is carried up to the full H.T. positive potential causing excessive beam current.

The same fault in the case of the cathode modulated circuit of Fig. 8.3(a) results in the tube *cathode* being carried up to H.T. positive potential, which causes the beam current to cut off.

For transistors, however, the situation is different. Should the transistor go open circuit, the collector moves to the full negative potential of the supply. With grid modulation the tube is safe because the grid is negative to the cathode, but with cathode modulation the cathode is carried negative with respect to the grid and high beam current flows. It is worth noting that the opposite fault, a short-circuited transistor, has the opposite effect and here the cathode modulation circuit has the advantage.

With the high level of reliability of transistors, however, it may perhaps be unnecessary to try to legislate against such failures.

4. *Cathode-to-heater voltage stress*

The higher the voltage between cathode and heater of the c.r.t. the greater is the chance of insulation breakdown. In valve receivers using series heater arrangements, the c.r.t. is placed at, or near, the 'earthy' end of the chain to give it maximum protection against the possibility of damage due to short circuits across the heater line. This places the heater at H.T. negative potential.

When grid modulation is used, since the cathode has to be higher in potential than the grid, it follows that the cathode voltage of the c.r.t. is at a higher potential than the video amplifier

anode (see Fig. 8.2(a)). Conversely, with cathode modulation, the c.r.t. cathode is at the *same* potential as the video-amplifier anode. The stress is therefore greater in the case of grid modulation.

In transistor receivers the c.r.t. derives its heater voltage from the normal low voltage supply used to feed the transistors. One side of this is tied to the chassis line which again places an H.T. potential between cathode and heater (see Figs 8.2(b) and 8.3(b)). In the case shown there is therefore a marginally greater safety factor with grid modulation, since the cathode is in this case nearer to chassis potential.

However, modern manufacturing techniques have minimised the likelihood of cathode-heater breakdown. It is unlikely therefore that this point will greatly influence the circuit designer in his choice of modulation method.

**Typical commercial circuits**

In the same way as was done for valve circuits in the previous chapter, some representative transistor circuits will be examined and attention drawn to points of interest.

The circuit of Fig. 8.7 is from the same receiver as that discussed at the end of Chapter 5 on vision detectors. It is recommended that the comments made in that chapter upon the detector

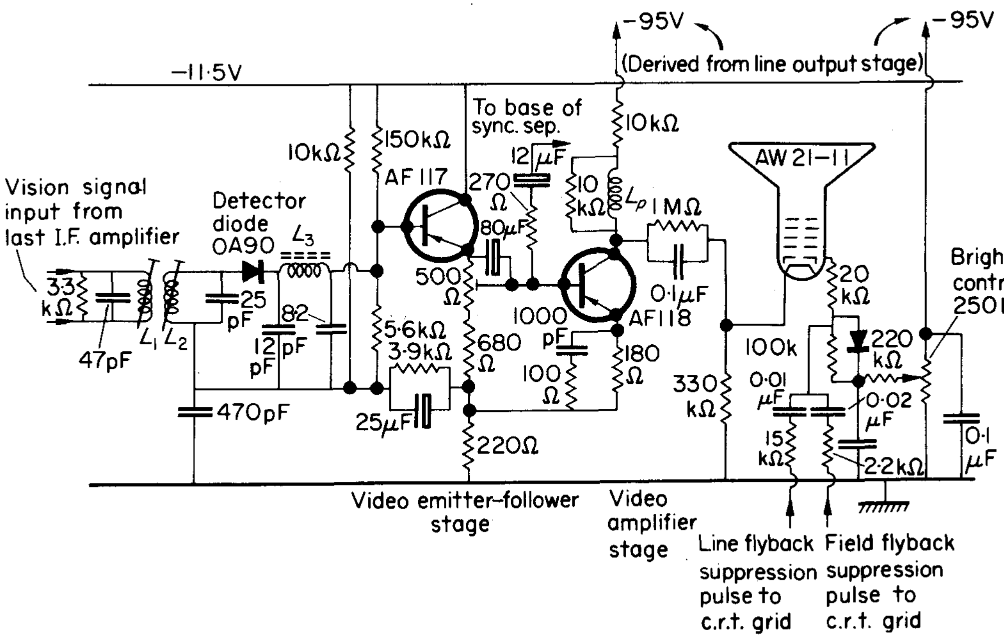



Fig. 8.7 Two-stage transistor video amplifier circuit in which a grounded-collector (i.e. emitter-follower) stage is inserted between the detector and a common-emitter amplifier
(*Note* that cathode modulation of the c.r.t. is used)

part of the circuit are re-read before proceeding further with what now follows. Having done so the circuit as a whole may then be looked at with reference to the following points:

(1) The c.r.t. is cathode modulated. A negative-going signal is needed at the collector of the video amplifier transistor, and this in turn calls for a positive-going input signal at its

base. This is provided from the emitter of the emitter-follower transistor. The latter is fed from the cathode of the vision detector diode, so that a positive-going signal is presented to the base and 'followed' at the emitter.

(2) A 500 Ω pre-set resistor in the emitter lead of the emitter-follower allows the base-bias current for the second transistor to be set at the optimum level. The part of this resistor which then falls in series with the base of the second transistor is by-passed by an 80 μF capacitor.

(3) The collector load circuit of the video amplifier includes a shunt-peaking coil ($L_p$) which is damped by a 10 kΩ resistor.

(4) A measure of emitter H.F. compensation is also employed. The 180 Ω resistor is not fully decoupled and negative feedback is developed at low and medium frequencies. At higher frequencies the effective emitter load diminishes as the reactance of the 1000 pF capacitor falls, and the negative feedback is removed to give a boost in gain at this end of the frequency range.

(5) A low frequency 'anti-flutter' circuit, consisting of a 0·1 μF capacitor and a 1 MΩ resistor, is included in the series coupling path to the c.r.t. cathode. Partial d.c. coupling is employed since the c.r.t. cathode returns to chassis via a 330 kΩ resistor. This, with the 1 MΩ resistor, forms a potential divider across the output terminals of the video amplifier.

(6) The auxiliary supply voltage for the video amplifier collector is seen to be −95 volts; this being derived from the line-output stage.

(7) Brightness control is achieved by returning the c.r.t. grid to a potential divider connected across the auxiliary supply voltage line.

(8) Both field and line flyback suppression pulses are applied to the c.r.t. grid. These must be negative-going pulses of sufficient amplitude to cut off the beam current during the flyback times. Some degree of differentiation of these negative pulses always takes place because they are fed through a capacitor and developed across a resistor in shunt with the c.r.t. grid. The diode conducts on the positive-going trailing edge of the pulses and short-circuits them to chassis, via a decoupling capacitor.

(9) The sync. separator is fed from the emitter of the emitter-follower stage. The positive-going video signal present here is of the required polarity for this purpose. (As explained earlier, the negative-going sync. pulses drive the p.n.p. sync. separator transistor into conduction.)

The circuit of Fig. 8.8 contrasts with Fig. 8.7 in that grid, instead of cathode, modulation of the c.r.t. is used. Note that:

(1) Brightness control is achieved by a potential divider connected between two points, one of which is at −68 V and the other +57 V relative to chassis. The resistor values are such that the cathode of the c.r.t. is positive with respect to the grid, by the necessary amount.

(2) A series peaking coil ($L_{py}$) is used in the collector circuit of the video amplifier between the collector and the video signal take-off point.

(3) In addition a shunt peaking coil ($L_{px}$) is inserted in series with the load resistor.

(4) Still further H.F. compensation is provided in the emitter circuit of this transistor. The 390 pF capacitor allows some negative feedback at the low and middle frequencies but removes it at the H.F. end.

(5) The sync. separator is also fed from the collector circuit of the video amplifier. This is

logical, since the transistor sync. separator, like the c.r.t. grid, requires a positive-going video signal.

(6) The vision detector faces its anode to the base of the emitter-follower stage. This stage receives a negative-going video signal input, and delivers a signal of the same polarity from its emitter to the base of the second transistor. This is of the sense needed to give a positive-going output at the collector of the second transistor.

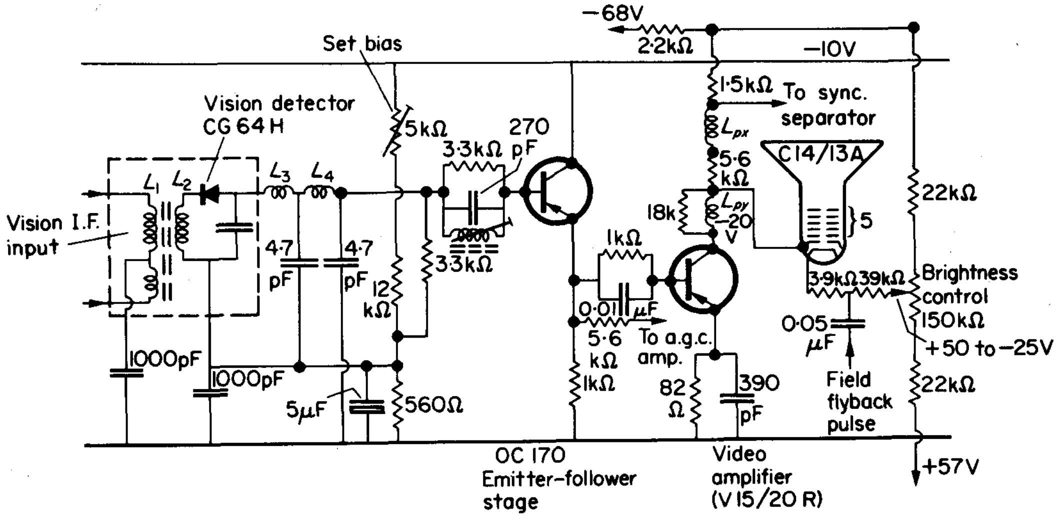



Fig. 8.8 Further example of a two-stage video amplifier using a common-collector (i.e. emitter-follower) feeding a common-emitter stage
(*Note* that grid modulation of the c.r.t. is used here)

(7) A parallel rejector circuit is inserted in series with the signal path from the detector to the emitter-follower. This is tuned to the inter-I.F. beat frequency which it prevents from reaching the video amplifier.

(8) The detector load resistor is of 3·3 kΩ and this is returned to a point on a potential divider to pick up the correct forward bias for the emitter follower stage. Adjustment of this bias is afforded by a pre-set 5 kΩ resistor.

The circuit of Fig. 8.9 shows a single-stage common-emitter video amplifier. Notice that:

(1) Grid modulation of the c.r.t. is employed. Hence the video amplifier must deliver a positive-going video signal at its collector, and is therefore fed at its base with a negative-going signal from the anode of the detector diode.

(2) The sync. separator, as would be expected, is also fed from the collector with its required positive-going signal. It should be noted that the resistive collector load is split into two parts by the 2·2 kΩ and 10 kΩ resistors. The sync. separator receives only that proportion of the total output signal amplitude which falls across the 2·2 kΩ resistor. The loading effect of the current-driven sync. separator is thus not shunting the complete collector circuit, and, moreover, a better match is provided to the sync. separator input impedance.

(3) A shunt peaking coil $L_p$ is employed to give H.F. compensation.

(4) Again, further H.F. compensation is provided at the emitter, since the emitter resistors are not fully decoupled. A pre-set resistor in the emitter circuit allows for an adjustment of the degree of negative feedback employed. It is called a *Quality Control.*

(5) $L_3$ and $L_4$ in the detector circuit form I.F. filter coils.

(6) $L_5$ which is in series with the detector load resistor of 3·3 kΩ gives a measure of H.F. compensation to the detector circuit. This is more necessary here than in a detector

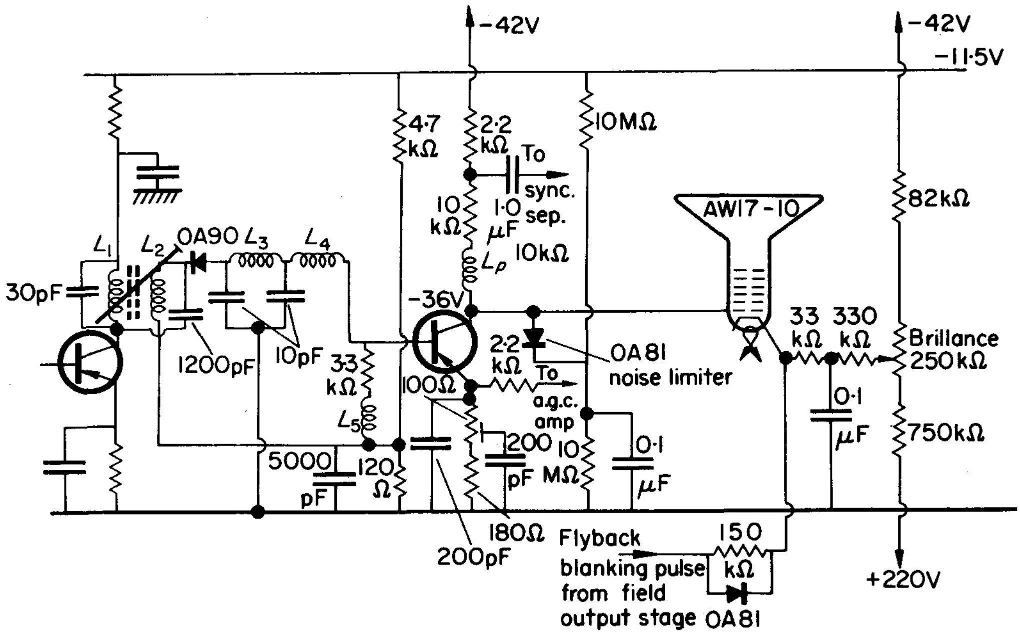



Fig. 8.9 Single-stage transistor video amplifier in a positive-modulation (405-line) receiver

which feeds an emitter-follower stage. The reason is that the input impedance to a common emitter stage is lower than that of an emitter-follower (i.e. common collector) stage and a greater degree of H.F. shunting is encountered.

Fig. 8.10 shows a further example of a two-stage video amplifier in which an emitter-follower feeds a common-emitter stage. Points to note are:

(1) Cathode modulation of the c.r.t. is used, so that a negative-going signal is needed at the collector of TR2. Since the circuit is designed for use in a positive modulation receiver, the vision detector faces its cathode to the emitter-follower base in order to give a positive-going input at this point.

(2) The diode load resistor of 4·7 kΩ returns to a potential divider network to pick up a suitable negative forward bias for the emitter-follower base. The 3·3 kΩ resistor is decoupled so that it does not form part of the diode load, but merely serves its d.c. purpose as part of the potential divider.

(3) The emitter-follower is coupled to the base of the common-emitter stage via a further I.F. filter $L_2C_2$, which assists the detector filter $L_1C_1$ in removing the I.F. component.

(4) The two transistors share an extra 470 Ω emitter-resistor. This resistor increases the load

resistance in the emitter circuit of the emitter-follower as far as the sync. separator is concerned, in order to provide a somewhat larger video signal input than is fed to the common emitter video-amplifier transistor. Notice that the latter receives its input only from across the 1 kΩ resistor.

(5) A combination of shunt and series compensation is employed in the collector circuit of the second stage. $L_{py}$, damped by a 27 kΩ resistor forms a series peaking coil, whilst $L_{px}$ is the shunt peaking coil.

(6) An auxiliary supply of −72 V is provided for the common emitter stage.

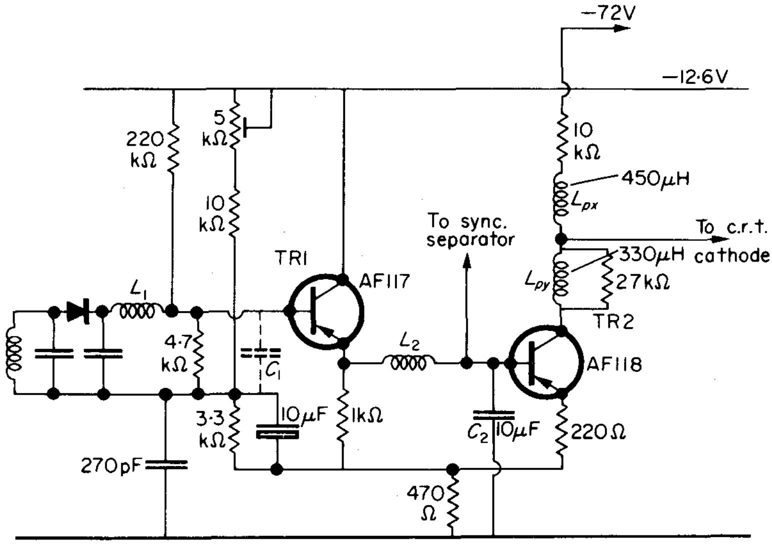



Fig. 8.10 Two-stage video amplifier in a positive-modulation receiver

A circuit for a negative modulation receiver is shown in Fig. 8.11. Note the following points:

(1) Cathode modulation of the c.r.t. is used. As in the previous circuit, in order to provide a negative-going video signal at the collector, the detector must feed a positive-going signal to the base of the emitter-follower stage. However, the diode must then face its anode to the base in order to give this polarity of output and not its cathode as when a positively-modulated signal is being handled.

(2) The emitter-follower is unusual in that it has 1 kΩ resistors in both the emitter and collector leads. The collector resistor is inserted to provide a load across which a suitable output may be obtained for driving the sync. separator. Thus this transistor, whilst behaving as an emitter-follower as far as the second video amplifier transistor is concerned, does in fact behave in the common emitter mode to feed the sync. separator. Clearly, the video-signal polarity at the collector is *negative-going*, as distinct from the emitter where a positive-going signal (following the base) appears. The negative-going output at the collector is required because in this particular circuit an n.p.n. transistor is used as the sync. separator. Such a transistor is fired into conduction by a positive-going voltage excursion. A negative-going video signal is therefore correct since the sync. pulses then represent the positive-going part of the wave-form. This compares with a normal valve sync. separator which also has to be fed with a negative-going video signal.

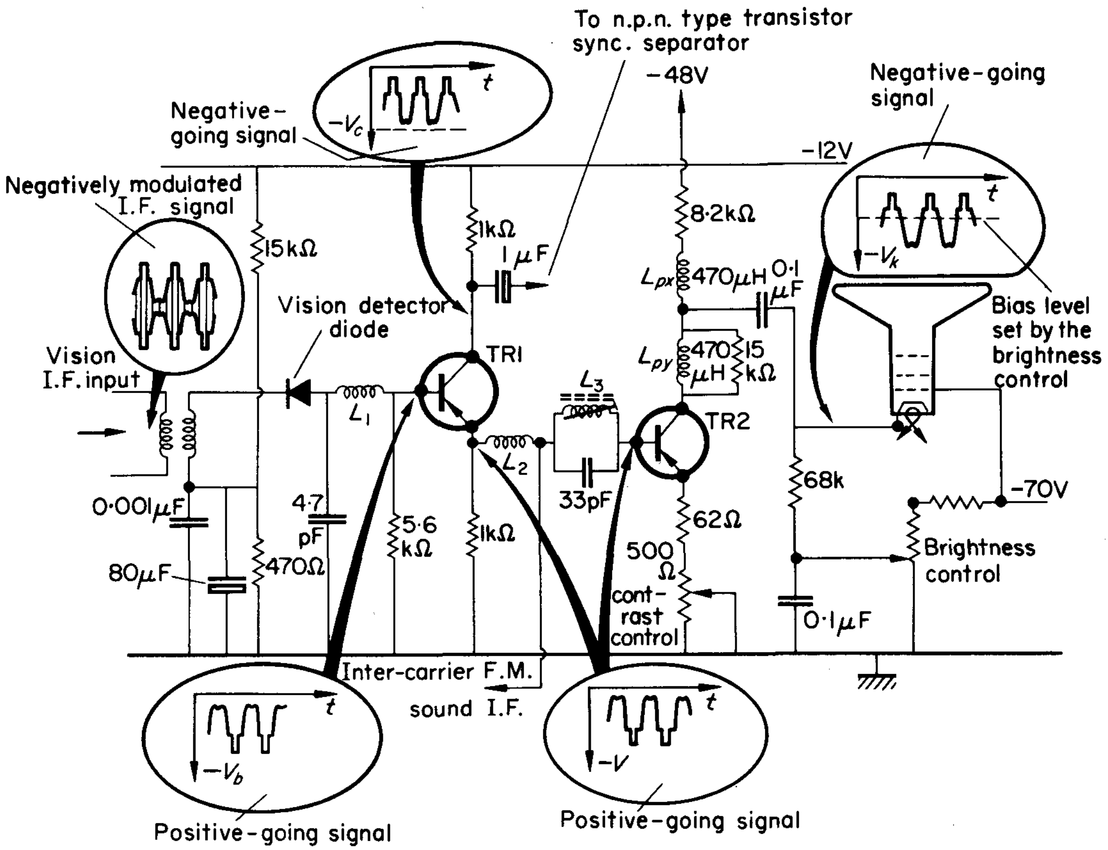



Fig. 8.11 Two-transistor video amplifier in a negative-modulation receiver

(3) As in the previous circuit, I.F. filter coils are present both between the detector and the emitter-follower, and between the latter and the second transistor. These I.F. filters are formed by $L_1$ plus stray capacitance in the detector circuit and $L_2$ plus stray capacitance between the two transistors.

(4) A parallel tuned circuit is also placed in series with the signal path to the base of the second transistor. This is there to take-off the inter-carrier F.M. sound I.F. signal used in the negatively-modulated system.

(5) Again, both series ($L_{py}$) and shunt ($L_{px}$) peaking coils are used to give the requisite degree of H.F. compensation in the collector circuit of the common-emitter stage.

(6) Contrast control is provided by means of a potentiometer in the emitter of the second transistor. It will be noticed that the emitter is not decoupled, and increasing the emitter resistor provides an increasing amount of negative feedback to the stage, with a corresponding reduction in gain.

(7) The c.r.t. is a.c. coupled via a 0·1 μF capacitor. Brightness control is achieved by varying the cathode potential relative to the grid. The grid is held constant at −70 V, and the cathode is clearly always positive with respect to the grid.

(8) The auxiliary supply to the video amplifier is −48 V in this circuit.

# 9

# Synchronising Pulse Separators: General Principles and Valve Circuits

## Purpose and required properties

This section of a television receiver is required to:

(a) Separate the synchronising pulses from the composite video signal. In effect this involves stripping off the picture information to leave a waveform consisting only of that part of the video signal which lies on the opposite side of suppression level to the picture information.

(b) Process the line and field pulses to make them suitable for their respective functions of synchronising the line and field timebase oscillators.

(c) Ensure good interlace by producing field pulses which are ideally of identical shape on odd and even fields. The timing of these pulses must be such that the *interval of time* between the *start* of the radiated field pulse sequence and the *triggering* of the field timebase is the same following odd and even fields.

(d) Be immune from the effects of normal changes in amplitude of the received vision signal. Despite the action of a.g.c. circuits the video signal reaching the sync. separator is bound to vary in amplitude, but the sync. pulses produced by it must not change their form because of this.

(e) Act in such a way that picture modulation changes do not affect the outgoing pulses in any way whatsoever.

(f) Be as immune as possible to the effects of interference 'noise'.

## Basic arrangements

Fig. 9.1 outlines the basic essentials.

The actual sync. separator slices off the sync. pulses from the video signal fed to it from the video amplifier. Following this, appropriate circuits, which may be simple C.R. networks or quite elaborate valve or transistor 'stages', segregate the line and field pulses and shape these to the desired form.

There is no fundamental difference between the circuit requirements for positive and negative modulation systems. The form of the video signal delivered by the video amplifier is the same in both cases. As pointed out in Chapter 3, however, the British 405-line positively modulated video signal does not include equalisation pulses before and after the field pulse sequence.

This makes for added difficulty in obtaining good interlace, as will be explained. Equalising pulses *are* present in the negatively modulated 625-line signal with the result that it is easier to achieve good interlace and hence the field pulse processing circuitry may be simpler. It will be

realised that this is not an inherent difference between positive and negative systems generally, but specifically a difference between the British 405-line and 625-line signals.

Noise, however, does affect the two systems differently and from the synchronising point of view it represents a greater problem with negative modulation systems. This may give rise to

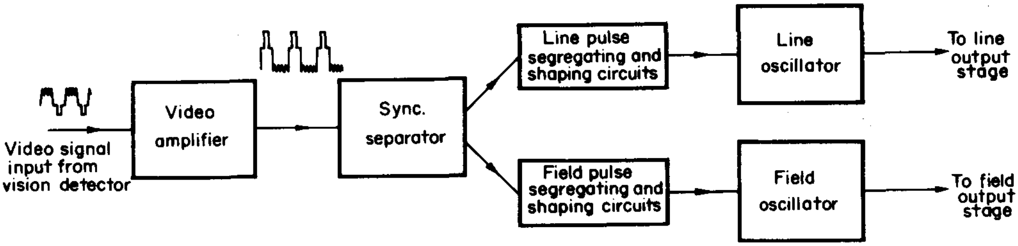



Fig. 9.1 Basic arrangement of synchronising pulse circuitry

some added complexity around the sync. separator. This matter will be examined in detail when the subjects of noise and interference—and in particular of noise-gated sync. separators—are studied.*

## Sync. separator circuits

The problem of taking off the sync. pulses from the video waveform is a comparatively simple one. A very widely used circuit is the pentode sync. separator. Fig. 9.2 shows a typical example from a dual standard 405/625 line receiver.

The lower the screen voltage of a pentode valve, the smaller is the negative grid bias voltage needed to cut off the anode current. By appropriate choice of anode and screen voltages the

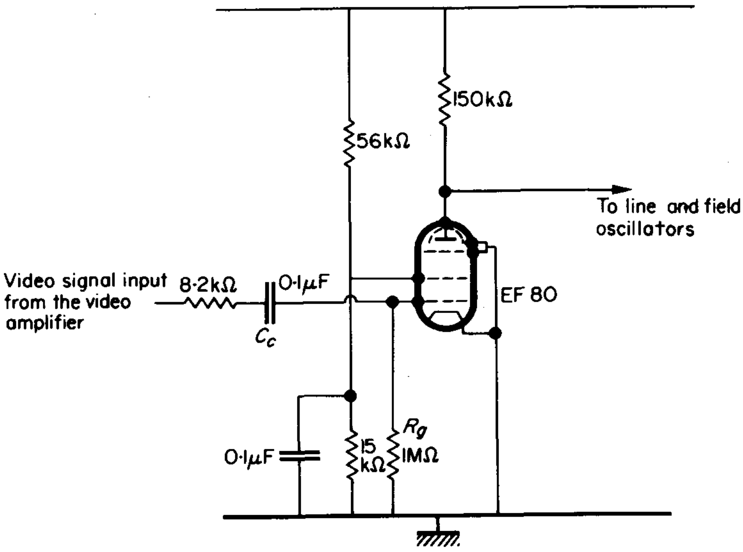



Fig. 9.2 Typical pentode sync. separator

* To be included in Part II of this work.

valve is arranged to operate with a short grid base of say 5 V width. The peak-to-peak value of the video signal voltage present at the output from the video amplifier is of an average order of 40 V to 60 V. The sync. pulses are therefore of the order 12 to 18 V amplitude.

The video signal is presented to the grid of the pentode with the sync. pulses as the most positive part of the waveform. In the quiescent state there is no bias on the valve. On the arrival of the signal the first few sync. pulses drive it into heavy grid current and the grid capacitor quickly charges up. The discharge current from the capacitor, which flows down through the grid leak resistor, drives the grid negative to cathode. The effect is that of an automatic negative bias so that the operating point sweeps back from $V_g=0$ to a point well beyond cut-off. After a few pulses the valve settles down into a steady state such that it is cut off completely except during the positive going sync. pulses. During the tips of these pulses it is driven into grid current which recharges the grid capacitor. There is a continuous flow of electrons from the capacitor through the grid resistor, and the p.d. set up across the resistor may be regarded as a steady negative bias whose amplitude remains constant all the while the applied signal amplitude is constant. The charge lost from the capacitor during the inter-pulse periods is replenished by the grid current which flows on the tips of the pulses.

The working conditions may be represented as shown in Fig. 9.3. It is as though the video signal from the anode of the video amplifier is coupled to the grid of a further amplifier stage which is working under 'Class C' conditions, i.e. is biased well beyond cut-off.

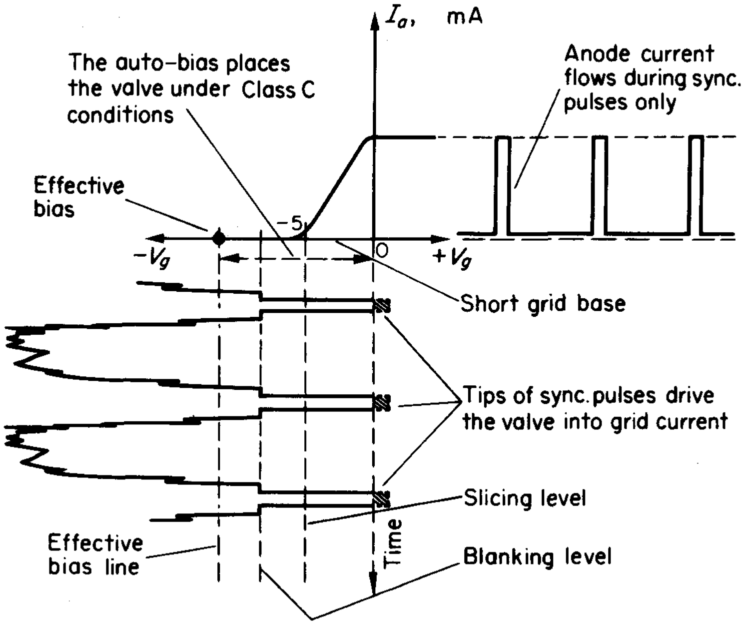



Fig. 9.3 Working conditions of a pentode sync. separator

The only parts of the video waveform extending under the grid base are the ends of the sync. pulses. These each cause a pulse of anode current which in turn give rise to corresponding negative-going voltage pulses at the anode. Between pulses the anode voltage rests at H.T. potential, since there is no volts drop across the anode load resistor. During the pulses the

anode voltage drops sharply to a much lower level. For example, if the H.T. voltage were 180 V, the anode potential may drop from +180 V to +80 V during pulses, giving negative-going sync. pulses at the anode of 100 V amplitude.

The actual minimum voltage reached, and hence the sync. pulse peak-to-peak amplitude, depends upon the value of anode load resistor and the peak anode current. A variety of pulse amplitudes from 25 V to 150 V or more, will be found from set to set. The designer chooses the operating conditions to give sharp-edged pulses of sufficient amplitude. A picture of the voltage waveform present at the sync. separator anode is shown in Figs 9.4(c) and (d). To clarify the

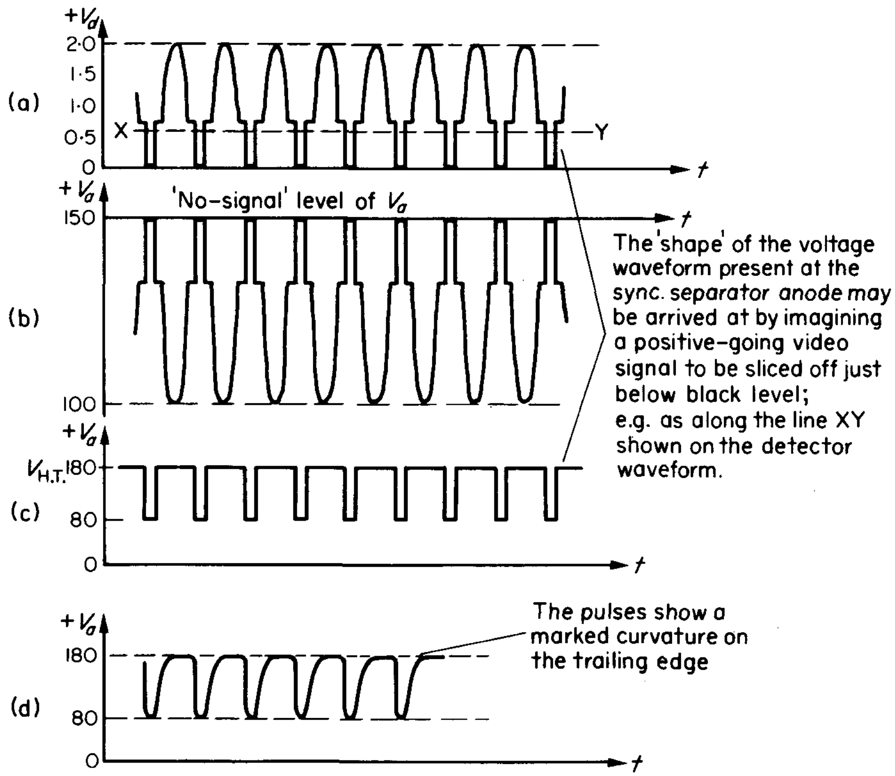



Fig. 9.4 Output voltage waveforms delivered by the vision detector, the video amplifier and the sync. separator
(a) Positive-going video signal as delivered by the vision detector to the grid of the video amplifier.
(b) Negative-going video signal as present at the video amplifier anode and passed to the sync. separator grid.
(c) Ideal voltage waveform present at the anode of the sync. separator.
(d) Practical output pulses showing the effect of shunt capacitance across the anode circuit.

matter further the corresponding video signal waveforms present at the detector and at the video amplifier anode are shown in Figs 9.4(a) and 9.4(b) respectively.

It should be mentioned that pulses can be too big as well as too small. In a given circuit, increasing the sync. pulse amplitude may result in the line oscillator triggering on half line pulses during the field sync. pulse sequence, and this causes an upset at the top of the picture until the line oscillator settles down once more to its correct frequency. The size of pulse needed will depend both upon the form of oscillator in use, and upon the point of application of the sync. pulses to the oscillator.

The presence of shunt capacitance in the anode circuit of the sync. separator modifies the pulse shape achieved. This manifests itself as a lack of squareness of the pulse tips and a general reduction in the steepness of the trailing edge. The rate at which the leading edge drops to the minimum pulse voltage level is delayed, and the climb-back to the 'resting' level at the end of the pulse is also delayed. The delay on the trailing edge is much greater because here the valve is switched off and plays no part in the effective anode circuit time-constant. However, it is the leading edge and not the trailing edge which is all important. Fig. 9.4(d) illustrates how the ideal square type pulse is modified by the presence of shunt capacitance.

Looking at the operation as a whole, of the complete video signal applied by the video amplifier to the sync. separator stage, the only part which gets through (i.e. is 'amplified') is the sync. pulse information. The sync. pulses straddle the grid base of the valve. The lower part of a sync. pulse is sliced off because it falls behind the cut-off point whilst the tip of the pulse is limited by grid current and anode current saturation. It is as though a section has been cut out of the sync. pulse, and for this reason the circuit is often described as a 'slicer'.

A pentode operating in this mode is very reliable. The valve is only lightly worked and the circuit functions efficiently with a wide range of input signal levels. The effective grid bias automatically adjusts itself to suit the signal amplitude.

The correct dynamic mutual characteristic is arranged by careful choice of anode load resistor and screen grid voltage.

As shown in Fig. 9.5, the load resistor is chosen so that the load line falls beneath the knee of the $I_a/V_a$ ($V_g=0$) curve. In the sketch shown the point of intersection is just above the $V_g=-1{\cdot}5$ V curve. This means that the valve 'bottoms' at a value of grid voltage just negative of $V_g=0$ V (i.e. between $V_g=-1$ V and $V_g=-2$ V). The student should study the significance of this statement most carefully. By projection to the left of the $I_a$ axis of the intersections of the load line with the $I_a/V_a$ curve, a series of corresponding values of $I_a$ and $V_g$ are obtained.

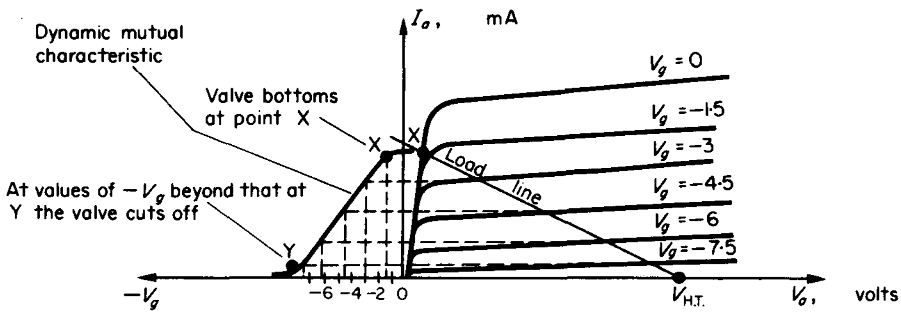



Fig. 9.5 Operating conditions for a pentode sync. separator

Plotting these gives the *dynamic mutual characteristic*, i.e. the relationship between $I_a$ and $V_g$ for this valve when used with the stated anode load resistor.

Suppose the negative bias voltage is reduced slowly from the cut-off point towards $V_g=0$ V. As the bias gets less the anode current steadily increases and there is a corresponding steady decrease in the value of anode voltage. Beyond point X however, decreasing $V_g$ makes no difference to the anode current. Thus at point X the anode voltage falls to a low value and stays constant despite further decreases of negative grid voltage.

There are two ways of changing the point at which the valve 'bottoms'. Clearly if the load

resistor is increased in value the load line will cut the $I_a/V_a$ curves at a lower point. Alternatively if the screen voltage is *increased* all the $I_a/V_a$ curves move bodily upwards. This again causes the valve to bottom at a more negative value of $V_g$ but, at the same time, the cut-off point moves to the left and the grid base broadens.

It is not surprising that a wide variety of combinations of anode and screen resistor values are found in pentode sync. separators. In general, increasing the value of the load resistor narrows the slicing range and makes for constancy of performance with a wide range of input-signal levels. Also the noise-bearing tips of pulses are clipped off more readily. On the other hand a point is reached beyond which further increases in value of the resistor cause the output pulses to be less sharp and well defined, with too pronounced a curvature on both the leading and trailing edges.

It should be noted that the effect of driving the valve into grid current on each sync. pulse tip is to clamp the sync. pulse tips to earth potential. Grid current charges the capacitor negatively to earth. The sync. pulses are prevented from excursing positively to earth, because as soon as they do so more grid current flows and the positive-going movement is negated by an increased negative potential across the capacitor. Looking outwards from the sync. separator grid the video signal now appears to be set entirely below earth potential. The sync. pulse tips are all at the same (i.e. earth) potential, despite wide variations in picture modulation content in the inter-pulse periods. (See Figs 9.2 and 9.3.)

The effect of the auto-bias action is thus to restore the d.c. component of the video signal, which has been lost due to the presence of the series d.c. blocking capacitor. If the pentode grid were momentarily disconnected from the input circuit, the video signal present across $R_g$ would be seen to be devoid of its d.c. component. Electrically it would set itself with equal areas on either side of the zero (earth) potential axis. A little thought will soon convince the student that if this signal were applied to a valve working as a normal amplifier, with a fixed bias instead of auto bias, then the output pulses would vary with picture content and with signal strength changes. It is vital that the d.c. nature of the video signal should be preserved. The sync. pulse tips must hence be tied to a fixed datum line and varying degrees of picture whiteness must correspond to levels offset by specific steps from black level. It is an easy matter to slice off the pulses when the signal is presented in this form. Of course, at the video amplifier anode, the video signal *is* so presented; the sync. pulse tips *do* correspond to a fixed datum line. It may be wondered why the signal is not d.c. coupled from here to the sync. separator. If everything would remain constant, this would be possible. However, it is inevitable that changes in d.c. conditions are bound to take place over a period of time, and this would make it very difficult to arrange for constantly efficient sync. separator action.

Either manual adjustments would be needed from time to time, or automatic re-setting circuitry fitted. Against this, the simple pentode sync. separator, a.c. coupled but with its own automatic and flexible d.c. restoration function, is an easy way out.

The diagrams of Fig. 9.6 illustrate this matter of the d.c. component, and of the clamping effect of the sync. separator auto-bias action.

In Fig. 9.6(a) the vision detector output voltage is shown for a bright line and a dark line at a given received signal strength, followed by the same two lines at a reduced signal strength level. Both in this diagram, and in the sketch of the corresponding waveforms at the video amplifier anode (Fig. 9.6(b)), all the signals are clamped with their sync. pulse tips at a constant level.

The dotted line shows a suitable slicing level. It is evident that uniform output pulses will result if either the grid or the anode waveforms are presented to a valve which is so biased that

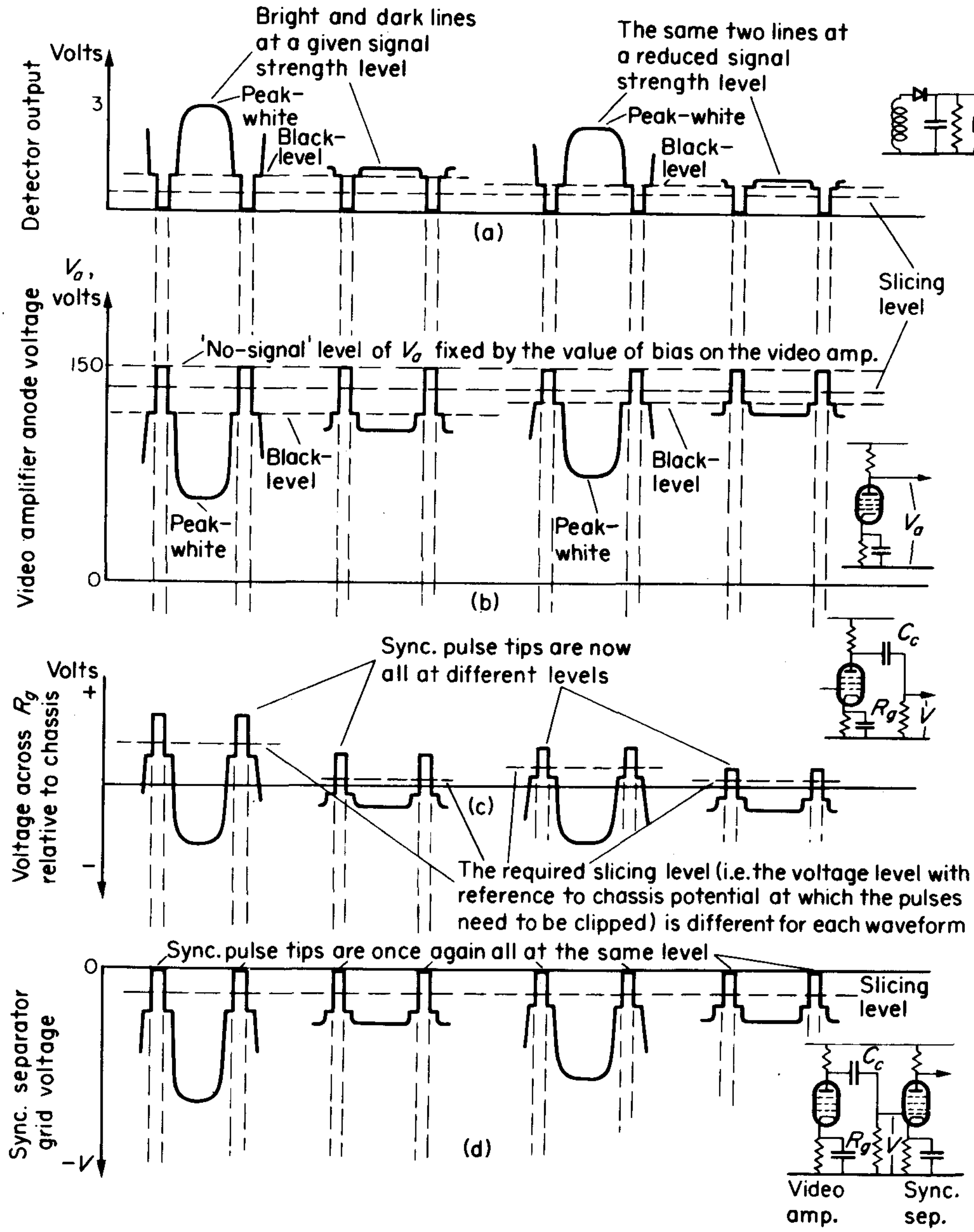



Fig. 9.6 The diagrams show video waveforms for bright and dark lines at two different signal strength levels. At the detector output and at the video amplifier anode, all sync. pulse tips rest at a constant level, i.e. the d.c. component is present. Uniform sync. pulses may be obtained by 'slicing' along the level shown

Fig. 9.6(c), however, shows the effect of a.c. coupling the video signal. There is now no possibility of obtaining constant amplitude sync. pulses by a simple slicing action.

In Fig. 9.6(d) the effect of d.c. restoration at the sync. separator grid is shown. Once again the sync. pulse tips are at a constant level and a slicing action is possible.

it only conducts during the part of the waveform falling between this level and the sync. pulse tips.

However, when the signals are a.c. coupled to the next stage via a coupling capacitor, the voltage waveforms across the resistor $R_g$ now take the forms shown in Fig. 9.6(c). Downward movements of anode voltage cause electrons to move down through $R_g$ making the top of it negative to chassis, whilst upward movements of anode voltage cause electrons to move up through $R_g$ driving the top positive to chassis. Since there must be equilibrium between the charge into and out of $C$ during a given period, it follows that the voltage waveform across $R_g$ will be such that the signal sets itself with equal areas either side of the mean (i.e. zero) voltage line. Neither black level nor any other level has any real permanent meaning now, since each waveform drawn shows different values for these.

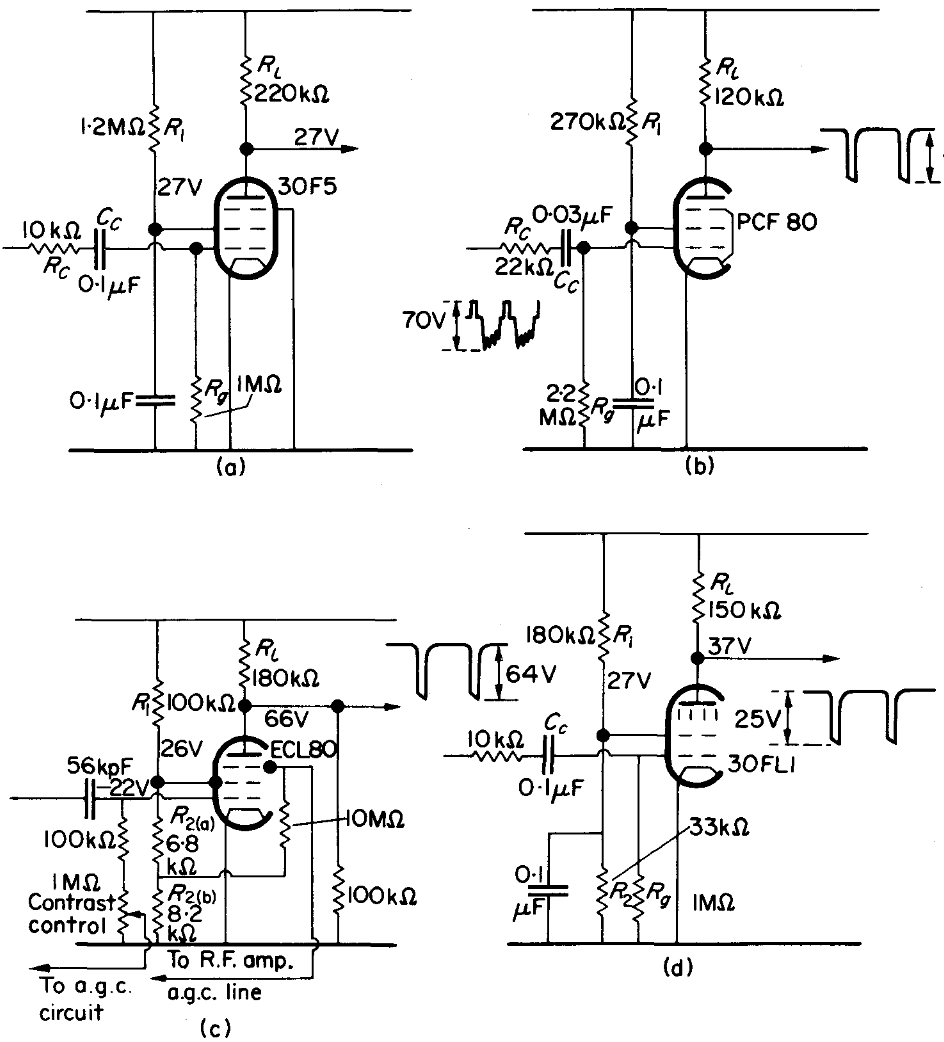



Fig. 9.7 Further examples of pentode sync. separator circuits

The necessary slicing level is seen to be at an entirely different voltage on all four waveforms. To produce uniform output pulses a sync. separator would have to have its bias adjusted to a different level for each line. Before attempting to produce sync. pulses it is essential that the d.c. component be restored. This is perceived to mean that the sync. pulse tips must be made to rest at a constant level. As already explained, this is achieved by driving the sync. separator into grid current. Fig. 9.6(d) shows that all sync. pulse tips now rest at chassis potential. The grid and cathode form a diode valve which functions in the same way as a d.c. restorer diode. The conditions at the sync. separator grid will be considered again when the subject of a.g.c is studied.

Some further examples of pentode sync. separator circuits which have appeared in domestic receivers are shown in Fig. 9.7.

In Fig. 9.7(a) and Fig. 9.7(b) the required low value of screen voltage is obtained by using a high value series screen-feed resistor. In Fig. 9.7(c), as in Fig. 9.7(d), a potential divider across the H.T. line is used for the same purpose. In this case it is the ratio of $R_1$ to $R_2$ which largely fixes the screen potential and $R_1$ is of lower value than in Fig. 9.7(a), where there is no additional bleed current to drop the voltage. Although a.g.c. circuits are not under discussion at the moment, it is worth noting in passing that $R_2$ in Fig. 9.7(c) is divided into two parts to provide across the lower resistor $R_{2(b)}$ a low positive bias voltage for application to the suppressor grid of the pentode. This grid then serves as the anode of a conductive *delay diode* which shorts the R.F. amplifier a.g.c. line to chassis until the incoming signal is strong enough to produce an a.g.c. voltage sufficiently negative to cut-off the suppressor grid. When this happens a negative a.g.c. potential is applied to the R.F. amplifier control grid and the gain of the stage is reduced.

The time constant of the grid capacitor and resistor is often of the order of 0·1 second.* A second resistor of the order of 10 kΩ is inserted in series with the d.c. blocking capacitor between the video amplifier anode and the sync. separator grid. This 'stands-off' the input capacity of the sync. separator from the video signal path to the tube. It serves to isolate the sync. separator from the picture signal path and enhances both the frequency response of the video amplifier and the efficient functioning of the sync. separator.

Triodes are sometimes used as sync. separators. The mode of operation is the same and the circuits function efficiently.

The pentode, however, does offer the designer a greater flexibility, and is usually preferred. By judicious choice of screen voltage and anode load resistor, the width of the grid base may be arranged for optimum performance in a given situation.

There is nothing highly critical in the design of sync. separators. Circuits such as that in Fig. 9.7(d) are capable not only of satisfactory operation for a wide variety of input signal strength levels but also handle both 405-line and 625-line video signals with equal dexterity, despite the quite significant difference in the time durations of the pulses being handled.

* E.g., in Fig. 9.7(a) the time constant is $C_cR_g$ seconds

$$= 0{\cdot}1 \times 10^{-6} \times 10^{6} = 0{\cdot}1 \text{ sec.}$$

In Fig. 9.7(b), $C_cR_g = 0{\cdot}03 \times 10^{-6} \times 2{\cdot}2 \times 10^{6} = 0{\cdot}066$ sec.

# 10

# Synchronising Pulse Separators: Transistor Circuits

*Basic circuit arrangements.* Transistors perform the same function as valves in sync. separator circuits and the mode of operation is closely analogous. Once again it is merely a question of so biasing the transistor that it is cut off during picture content and only conducts on sync. pulses. As with a valve, it is necessary to make the bias self-adjusting so that the operating point shifts with changes in incoming signal strength.

Fig. 10.1(a) shows a typical circuit. The transistor is a p.n.p. type and hence conducts when the base is driven negative with respect to the emitter and cuts off when the base is positive to the emitter.

With the valve circuit, positive-going sync. pulses cause grid current which charges up the grid capacitor to produce the necessary auto-bias. With the p.n.p. transistor, negative-going sync. pulses are needed to perform the same function, so that the video signal input to the base has to be positive-going. The base emitter junction then rectifies the input signal, conducting on the sync. pulses which are the most negative part of the input waveform. When first switched on, successive sync. pulses cause the series capacitor to charge up to a voltage approximating to the peak value of the video signal. Thereafter, with a properly chosen time constant, the capacitor discharges slightly between pulses, and the amount of charge lost is replaced by base current during the tips of the sync. pulses.

If there were no leakage current to the collector there would be no volts drop across the collector load resistor between pulses, and the collector would rest at the applied battery potential. Since the current during sync. pulses is made to 'bottom' the transistor, the output sync. pulse amplitude would then be almost equal to the battery supply voltage; in this case 12 V. Some leakage current is inevitable, however, so that the maximum available pulse amplitude is somewhat less than this.

To preserve the all-important steep-fronted leading edge of the sync. pulse a transistor capable of a fast switching action is chosen, and for optimum performance this quality must be combined with low leakage current and an adequate current gain. It is instructive to pause to note that the electron current through a valve may be switched on or off instantaneously by application of rectangular pulses to the grid. A transistor behaves differently and it does not switch off as fast as it switches on. This is due to the phenomenon known as 'hole storage' as a result of which the cessation of hole-injection into the base region (i.e. the switching off of the input base-emitter current) does not immediately stop the flow of holes from the base to the collector.

Once the transistor has been conducting, a greater than normal number of holes exists in the base region and these keep the collector current flowing for a brief period after the removal of the forward bias on the base-emitter junction. The effect is similar to the 'soak charge' in a

capacitor dielectric, as a result of which it is possible to produce a second spark from a capacitor which has a few moments previously been 'discharged' by a temporary short circuit across its plates. Some transistors have a more efficient switching action than others, so that this is a quality which is studied in the selection of transistors which are to work as sync. separators.

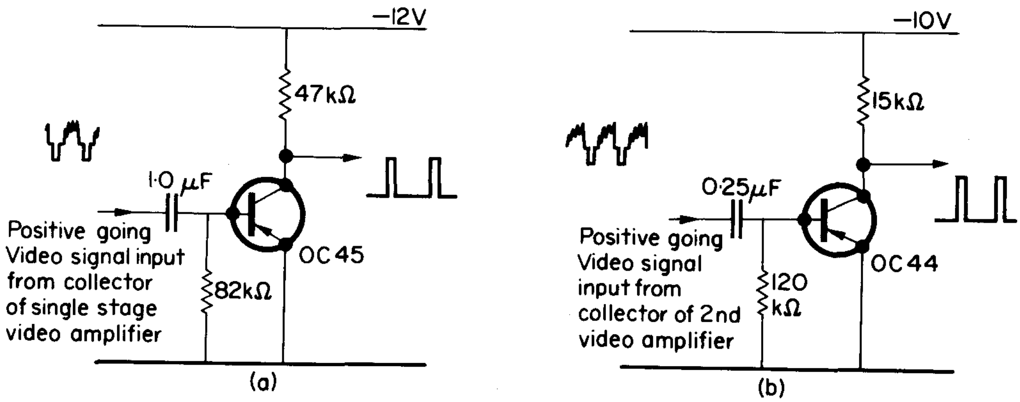



Fig. 10.1(a) Self-biased transistor sync. separator

Fig. 10.1(b) A further example of a self-biased transistor sync. separator

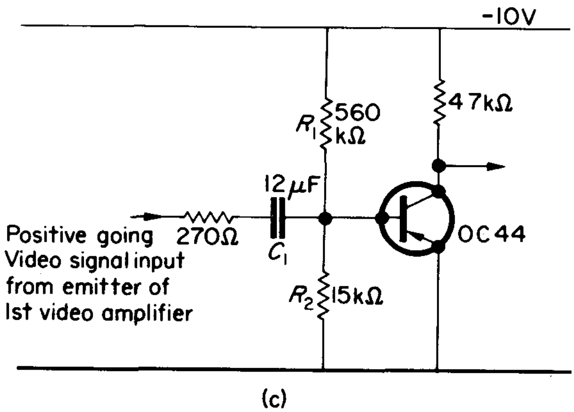



Fig. 10.1(c) Transistor sync. separator employing combined fixed-bias and auto-bias

A transistor parameter which gives an indication of its switching-speed capability is $f_{co}$, the cut-off frequency. This is defined as the frequency at which the current gain of a transistor falls by 3 dB and it is obvious that a transistor with a low cut-off frequency cannot behave as a fast-acting switch. To handle pulses of the order of duration met in television receivers (e.g. in the British case, 6·4 μs and 10 μs for the line sync. pulses of 625 and 405 signals respectively), a cut-off frequency in excess of 2 Mc/s is called for.

As the sync. separator transistor is switched off for most of the time, it dissipates very little power and hence a small R.F. transistor (e.g. of the OC44 type) has the qualities needed for this circuit.

The discharge time-constant of the input circuit of Fig. 10.1(a) is

$$(1 \times 10^{-6} \times 82 \times 10^{3}) \text{ seconds} = 82 \text{ ms}$$

whilst that of Fig. 10.1(b) is

$$(0{\cdot}25 \times 10^{-6} \times 120 \times 10^{3}) \text{ seconds} = 30 \text{ ms}$$

The former circuit has a collector load of 47 kΩ and the latter 15 kΩ. In general a higher value load resistor reduces the peak drive current needed to bottom the transistor but leads to a broadening of the output pulse. This is due to the hole-storage referred to. When the transistor base current is cut off at the end of the pulse, the holes stored in the base take longer to dispel if the collector load resistor is increased.

There is a marked difference in the signal voltage amplitudes necessary to drive valve and transistor sync. separators. The valve circuit is fed with the fully amplified video signal present at the anode of the video amplifier, which is usually of the order of 50 V or more. A transistor on the other hand needs only a sync. pulse input amplitude of the order of 200 mV to 400 mV. If 300 mV is taken as the required sync. pulse amplitude, then taking the sync. pulse as 30% of the peak-to-peak video signal voltage, it follows that a minimum video signal input is needed, of amplitude:—

$$\left(\frac{300}{30} \times 100\right) \text{mV} = 1000 \text{ mV} = 1 \text{ V}$$

To sum up it is seen that a transistor sync. separator needs a video signal input of the order of 1 V peak-to-peak and that this must be a positive-going signal (i.e. negative-going sync. pulses) for a p.n.p. transistor or a negative-going signal (i.e. positive-going sync. pulses) for an n.p.n. transistor. An input signal of this order of amplitude then gives rise to output pulses whose amplitude approaches the value of the supply voltage; e.g. 12 V for Fig. 10.1(a) or 10 V for Fig. 10.1(b).

The circuit of Fig. 10.1(c) draws attention to a different biasing arrangement. Here a combination of fixed and auto-bias is employed. The resistors $R_1$ and $R_2$ form a potential divider across the supply voltage, and under no-signal conditions this provides a forward bias of about 250 mV to the base-emitter junction. On application of the positive-going video signal, the negative-going sync. pulses cause sharp increases of base emitter current and $C_1$ charges up. The discharge current of electrons up through $R_2$ to $C_1$ then drives the base positive to the emitter, so biasing the base-emitter junction well beyond cut-off during the picture content part of each line.

As with the previous circuits the partial discharge of $C_1$ between pulses is replaced by base current during the tips of the pulses. The fixed forward bias ensures that the tips of the pulses drive the transistor well into the bottoming condition to give good clean output pulses.

*Polarity of pulses.* It is instructive to spend a little time studying the polarity of current and voltage pulses in transistor sync. separators. As a starting point, the diagrams of Fig. 10.2 allow a comparison of valve and transistor circuits.

In describing the passage of signals through valve circuitry, it is customary to use the chassis as the reference level and to refer to signal conditions at various points by specifying the directions relative to chassis of the variations in signal potentials at these points. It is the H.T. supply to a valve circuit which dictates the polarity of the mean (or no-signal) potential at an anode. This is positive with respect to chassis, and in the case of the sync. separator it may be assumed that the anode rests at the full H.T. potential in between sync. pulses. The arrival of

a positive-going sync. pulse at the grid causes the anode voltage to fall and we speak of *negative-going* output pulses. This is quite logical and meaningful since it explains clearly the movement of potential at the anode, relative to the chassis. The point is illustrated on Fig. 10.2(a). There is a further subtle factor which, though not electrical in character, none the less affects our

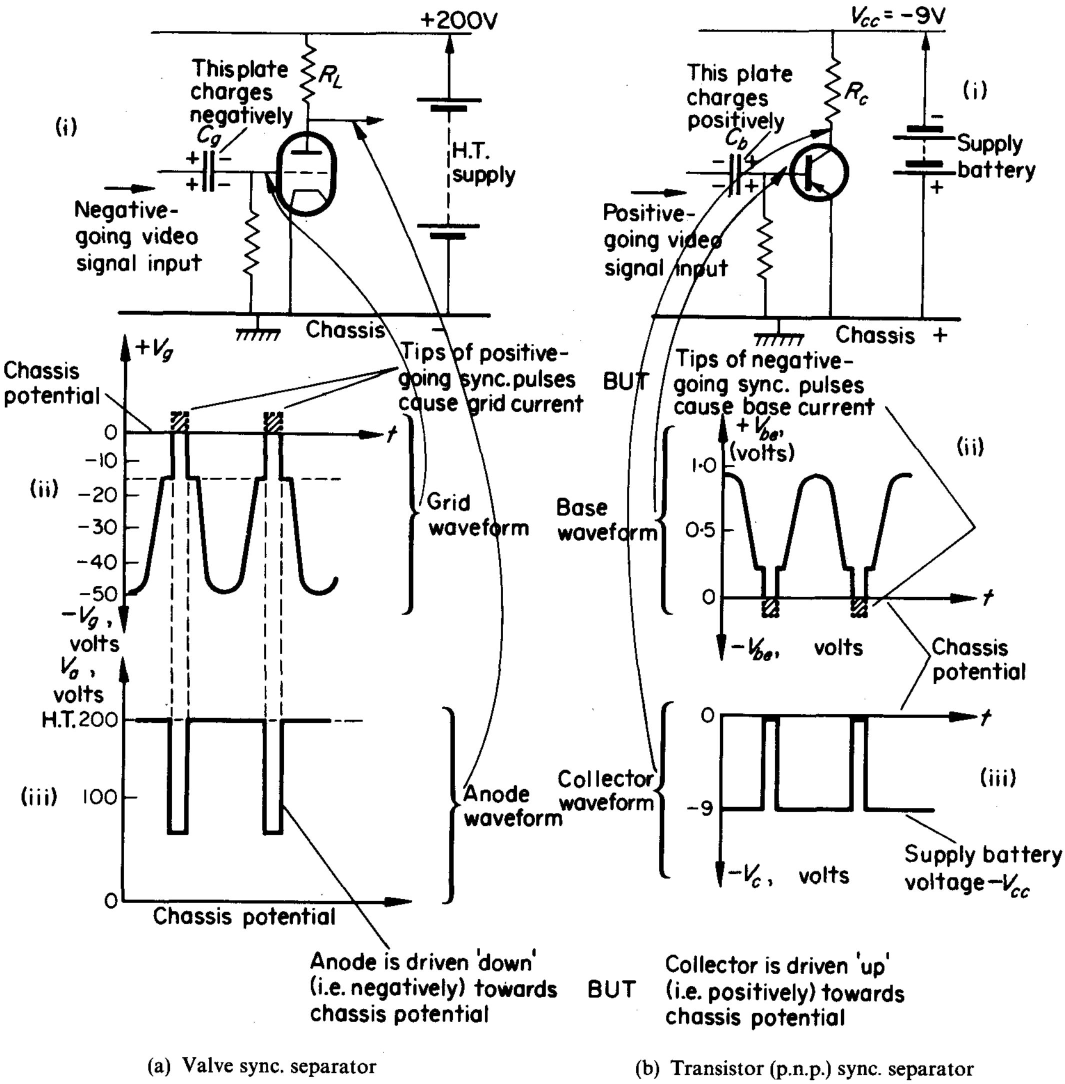



Fig. 10.2 Signal polarities in valve and p.n.p. transistor sync. separators

thinking. A valve circuit diagram shows the chassis at the *bottom* of the circuit, and it appears natural to speak of the anode potential as being driven *downwards* towards chassis potential or *upwards* towards the H.T. line potential. Fortunately in valve circuitry this is right electrically as well as appearing to make gravitational sense!

The p.n.p. transistor upsets our reasoning. To begin with the upper line in the diagram,

corresponding to the H.T. line, is negative with respect to the lower line. The latter is often, but not always, the chassis line. In the diagram of Fig. 10.2(b) it *is* the chassis; i.e. the battery positive is connected to chassis. The emitter is tied to chassis and the transistor passes collector current when the base is driven negative to chassis. The shaded tips of the sync. pulses show the moments of conduction, and for the rest of the waveform the base is positive to chassis and no collector current flows. When no current flows, the collector rests at the potential of the upper line which is 9 V negative to chassis. The pulses of collector current result in a volts-drop across the collector resistor so that the potential here moves towards the chassis potential, and in fact, if the transistor bottoms, it does almost reach chassis potential. There is a temptation to say that the collector potential is driven downwards towards the chassis potential. This is of course wrong, because the chassis is positive and to be correct we should have speak of the collector as being driven *upwards* towards chassis potential. This tends to throw us out of balance and takes a little getting used to. The waveform sketches on the diagram show the sync. pulses at the collector as positive-going pulses which take the collector *up* to the chassis potential.

The matter is still further complicated by the fact that it is the *upper*, or battery *negative* line, which is connected to chassis in some circuits. The potential of the collector is then equal to chassis potential when the transistor is nonconductive, but moves positive to chassis when the sync. pulse causes collector current to flow.

It has been said that connecting the chassis to battery negative confers the advantage of making all voltage readings positive to chassis as they are in valve circuits. In practice this is not an advantage at all. A great many circuits have battery positive connected to chassis and a technician gets used to checking to make sure that the necessary negative forward bias is present on a p.n.p. transistor amplifying stage by measuring the base and emitter voltages relative to chassis; e.g. a base reading of $-1{\cdot}25$ V and emitter reading of $-1{\cdot}1$ V shows a forward bias of $-(1{\cdot}25-1{\cdot}1)=-0{\cdot}15$ V. Should the chassis be connected to battery negative, however, and the testmeter positive lead habitually clipped to chassis, the readings at the base and emitter drive the pointer backwards. The first reaction is to change the leads over, in which case, assuming a battery voltage of 9 V in the example discussed, the base reads $+7{\cdot}75$ V (i.e. $9-1{\cdot}25$) and the emitter 7·9 V (i.e. $9-1{\cdot}1$ V). Of course, a little thought then confirms that the base is at a potential of $-0{\cdot}15$ to the emitter, but no one could pretend that this connection is an aid to understanding. In circuits such as this, some manufacturers wisely instruct the engineer to take all readings with the meter positive lead connected to battery positive. In this event the chassis does not feature in the readings as a reference line at all. Readings are always recognisable if taken with respect to the supply line to which the transistor emitters are returned, and if this is done it does not matter where the chassis features in the diagram.

Clearly the answer is to get used to thinking in terms of transistors when handling transistor circuits and vice versa with valves; just as a true linguist thinks in the language he is speaking in and does not go through the added complication of framing his thoughts in his mother tongue and mentally translating them. To achieve this mental dexterity it is important to be as familiar with transistor characteristics as with valves.

*Transistor characteristics.* In discussing valve sync. separators a dynamic mutual characteristic was studied (see Fig. 9.5). The corresponding input/output, or 'transfer', characteristic for transistors is the $I_c/I_b$ curve.

As with the valve a distinction must be drawn between the static $I_c/I_b$ curve which is plotted with $V_c$ held constant, and the dynamic curve which takes account of the fact that when $I_c$ varies, $V_c$ also changes because of the changing volts drop across the collector load resistor.

In Fig. 10.3 a load line is drawn across a family of static $I_c/V_c$ curves.* From the intersections of the load line and the static curves, a dynamic $I_c/I_b$ transfer characteristic is then drawn. In the example shown a small R.F. transistor is employed working on a 10 V supply with a 5kΩ collector load resistor. The transistor bottoms (i.e. $V_c$ falls to nearly zero) when $I_b = 30$ μA.

Sync. separator circuits often employ fairly high values of collector load resistor, with the result that the transistor bottoms at surprisingly low values of base input current. For example, using a 20 kΩ load resistor and a 10 V supply, $V_c$ falls to zero when

$$I_c = \left(\frac{10 \times 10^3}{20{,}000}\right) \text{mA} = 0{\cdot}5 \text{ mA}$$

The base current necessary to set up such a low value of collector current may only be of the order 10 μA to 15 μA. The implication is that a very small input signal is sufficient to cause the

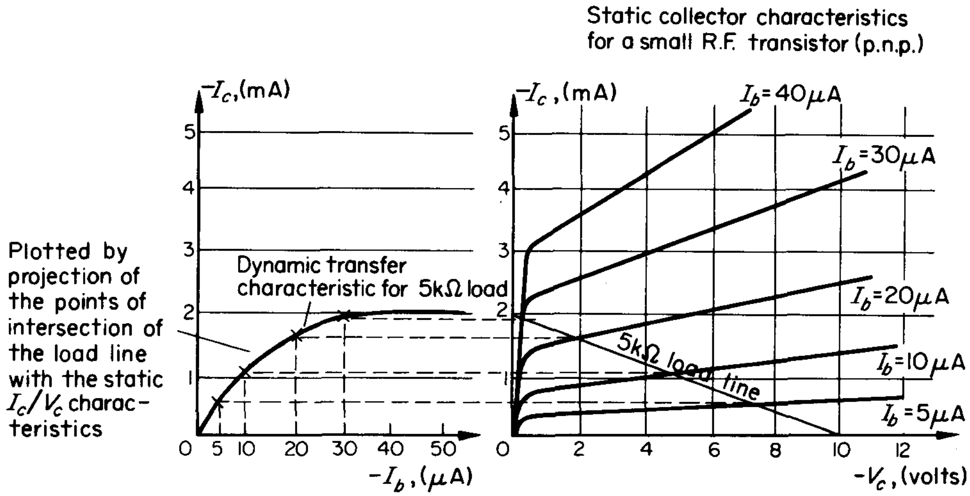



Fig. 10.3 Showing how a dynamic transfer characteristic is obtained by plotting a load-line on a family of $I_c/V_c$ curves and noting the intersections

collector voltage to drop to zero from its resting (or 'inter-pulse') level. The latter is almost equal to the battery voltage if the leakage current is low. Full amplitude output sync. pulses are hence developed for a wide range of input signal strength levels.

The method of obtaining a transistor dynamic transfer characteristic is seen to be the same as for the valve. An important difference occurs, however, between the use of valve ($I_a/V_g$) and transistor ($I_c/I_b$) transfer curves as pictorial devices to illustrate the behaviour of signal-handling circuitry. In Fig. 9.3 the input signal to the valve is plotted against the $V_g$ axis, and is

* The load line shows the true dynamic relationship between the collector current $i_c$ and the collector voltage $V_c$ when a given load resistor is present in the collector circuit. This numerical relationship is given by the expression $V_c = \{V_{cc} - I_c R_l\}$ and the load line is simply a plot of this 'straight-line' law. Since the graph is a straight line, only two points are needed to plot it and these are given by the two limiting values as follows:

When $I_c = 0$; $V_c = V_{cc}$ (In the example shown this is $-10$ V)

When $V_c = 0$; $I_c = \dfrac{V_{cc}}{R_l}$ $\left\{\text{In the example shown this is given by } I_c = \dfrac{10}{5000}\text{ A} = 2\text{ mA}\right\}$

fully represented in terms of a voltage having the dimensions shown on the $x$-axis of the valve's $I_a/V_g$ curve. This is perfectly realistic, because the input signal *does* indeed exist as a voltage of the magnitude shown. The only reservation is that the part of an input signal which drives the grid positive will be modified in shape by grid current. This is recognised on most diagrams by shading, or just dotting in, that part of the signal which swings past the origin into the $+V_g$ region. It should be noted that the part of the input signal which does not fall below the valve's $I_a/V_g$ curve, i.e. is to the left of the $V_g$ cut-off point, still exists as a voltage and the instantaneous magnitude of it is truly represented by the $-V_g$ voltage scale.

It is not correct, however, to use the transistor $I_c/I_b$ curve in the same way. The temptation may be to plot a scale showing $+I_b$ to the left of the origin and $-I_b$ to the right. Quite clearly it is meaningless to draw an input signal labelled $\pm I_b$ beneath the $I_c/I_b$ curve. It is true that the part of the signal falling beneath the curve itself (i.e. the part extending into the $-I_b$ region) does *exist* as a base current, but the rest of the signal gives rise to no base current at all. The only time it is permissible to draw a signal beneath an $I_c/I_b$ curve is when the signal so shown is *in fact* the *current* input waveform to the transistor. For example in a Class A amplifier, if it is known that a sine-wave input current is being superimposed upon the steady forward bias current of a transistor, then the input signal is correctly represented by a sine wave drawn beneath the $I_c/I_b$ curve, with the forward bias level as the sine wave's axis.

With an operating mode such as that in use with sync. separators (which is really a Class C condition), the only way in which the input signal applied to the transistor may be fully represented is to draw it as a *voltage* input to the base-emitter junction. Occasionally transistor manufacturers publish curves of collector current $I_c$ against base-emitter voltage $V_b$. In Fig. 10.4 such a curve is drawn and the video input signal is shown disposed beneath it as it would exist in a sync. separator circuit. This is not a true picture of the circuit behaviour, however, since the $I_c$ curve sketched is a static characteristic for a fixed value of $V_c$ and does not represent the dynamic condition.

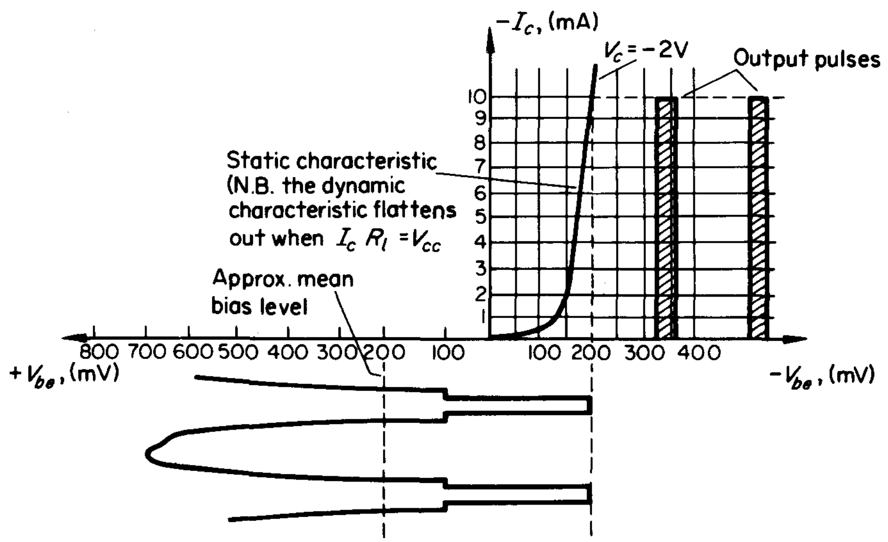



Fig. 10.4 Showing a p.n.p. transistor $I_c/V_{be}$ characteristic with the video signal input fully represented as an input voltage

More usually the curves published for the common emitter mode are the output characteristic ($I_c/V_c$), the transfer characteristic ($I_c/I_b$) and the input characteristic which shows the base input current $I_b$ plotted against the base-emitter voltage $V_b$. In order to demonstrate the approximate behaviour of a transistor in a sync. separator circuit these may be used as follows:

(a) Plot a load line across the $I_c/V_c$ curves for the selected collector load resistor (Fig. 10.3).
(b) Plot the dynamic $I_c/I_b$ curve from the intersections of load line with the $I_c$ curves (Fig. 10.3 and Fig. 10.5(b)).
(c) Draw the video signal input voltage beneath the $I_b/V_b$ input characteristic. Here the signal must be drawn with equal areas on each side of the effective bias line. Having done this the pulses of base current $I_b$ which result from the sync. pulses which fall beneath the $I_b/V_b$ curve, may now be drawn (Fig. 10.5(a)).
(d) Having arrived at the true dimensions of the input ($I_b$) current pulses these may now be drawn beneath the dynamic transfer characteristic obtained under (b), and the nature of the collector current is then shown (Fig. 10.5(b)).
(e) If desired the collector output voltage pulses may now be deduced by taking the product $I_c . R_c$, where $R_c$ is the collector load resistor. These pulses may then be depicted in the way shown in Fig. 10.2(b)(iii).

It should be noted that the shape of the dynamic characteristic is similar to that of a pentode and results in 'topping and tailing' the input sync. pulses. By cutting off the noise-bearing pulse tips, 'clean' output pulses are produced.

Manipulating published characteristics in this way is very well worthwhile since it leads to a better mental picture of what is happening in transistor circuitry. An understanding of the relationships which exist between the input and output currents and voltages of the transistor itself is a pre-requisite to an understanding of the way in which a given circuit containing transistors manages to achieve its desired purpose. It should be stressed, however, that the method outlined above leads only to an approximate representation of the signal and pulse dimensions which exist in the functioning transistor sync. separator circuit. To begin with the curve quoted under (c) is itself a static curve which shows how $I_b$ varies against $V_b$ for a *constant* value of $V_c$. A more accurate picture would require the use of a dynamic input characteristic which would reflect the variation in $V_c$ which *must* result as $I_b$ (hence $I_c$) varies, since the volts drop across the collector load resistor changes with each change of $I_c$. Other influences which bear on the final actual performance are transistor and associated circuit capacitances together with the phenomenom of hole storage. The transistor cannot switch on and off instantaneously, nor can the associated circuitry handle instantaneous changes in voltages and currents without introducing some delay. These factors lead to the sync. pulses showing curvature, particularly on the trailing edges.

Before leaving the subject of transistor characteristics, it is worth looking at one other point.

To avoid unnecessary complications, it has become customary to plot transistor characteristics, both p.n.p. and n.p.n., the same 'way up' as valve curves, irrespective of the true sign of the voltages and currents which are being portrayed. This is good practice, but none the less it gives added insight into the question of signal polarities if this matter is explored a little further. One needs to be *aware* of true polarities, even though the curves do not always show them.

To be strictly accurate mathematically, and to show their true relationship to valve characteristics, it would be necessary to plot the curves for p.n.p. transistors 'upside-down' as shown in Fig. 10.6(b). The only merit of doing this is that when they are so drawn the true polarities of

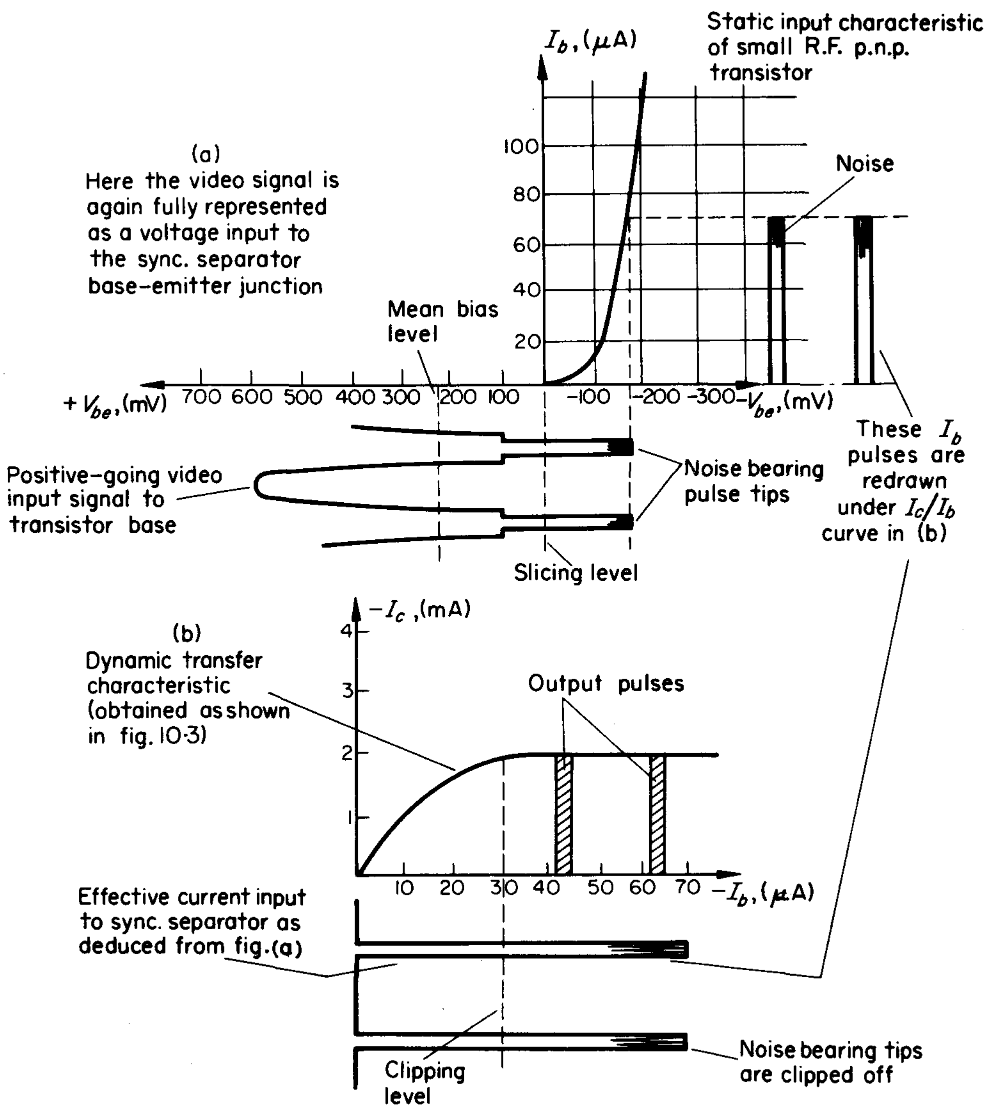



Fig. 10.5 Showing the input voltage waveform and the resulting base and collector current pulses

the input signals applied to the transistor and the output signals delivered by it are clearly seen, and a comparison with a corresponding valve circuit may more readily be obtained.

Thus in the case of a valve sync. separator, a negative-going video signal input is required so that the valve is driven into conduction by the 'positive-going' sync. pulses.

In the case of a p.n.p. transistor, a positive-going video signal is needed so that the transistor is driven into conduction by the negative-going sync. pulses. The terms positive- and negative-going are here used with their normal meaning; i.e. by positive-going we infer that increasing brightness drives the voltage level positive with respect to black-level. The diagrams of Figs 10.6(a) and 10.6(b) show the corresponding cases.

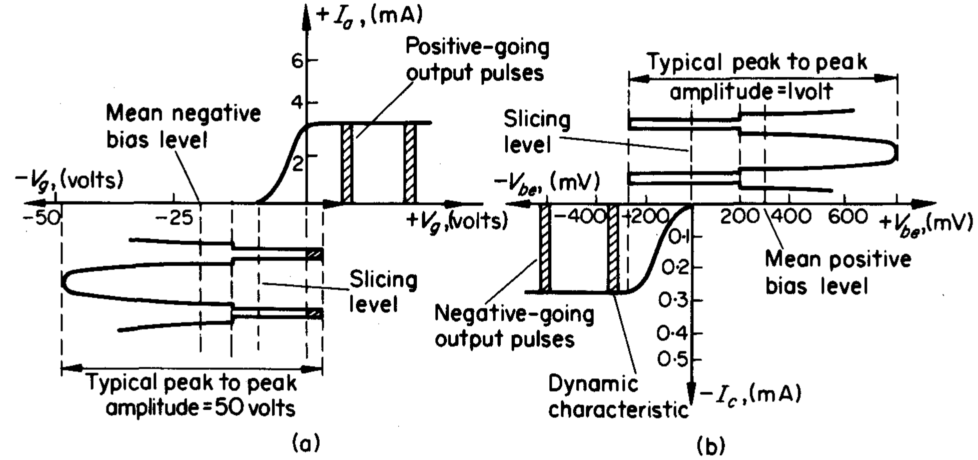



Fig. 10.6 Comparison of valve and transistor sync. separator operating conditions
(a) Valve—A negative-going video signal input of 50 V amplitude gives rise to positive output current pulses, with corresponding negative-going voltage pulses at the anode.
(b) P.n.p. transistor—A positive-going video signal input of 1 V amplitude gives rise to negative output current pulses with corresponding positive-going voltage pulses at the collector.

With n.p.n. transistor sync. separators, a negative-going video signal is necessary, since in this case it is positive-going sync. pulses which are needed to drive the base-emitter junction into conduction.

As an exercise in studying signal polarities the circuit of Fig. 10.7 is instructive. Here a p.n.p. transistor sync. separator is used in a hybrid receiver.

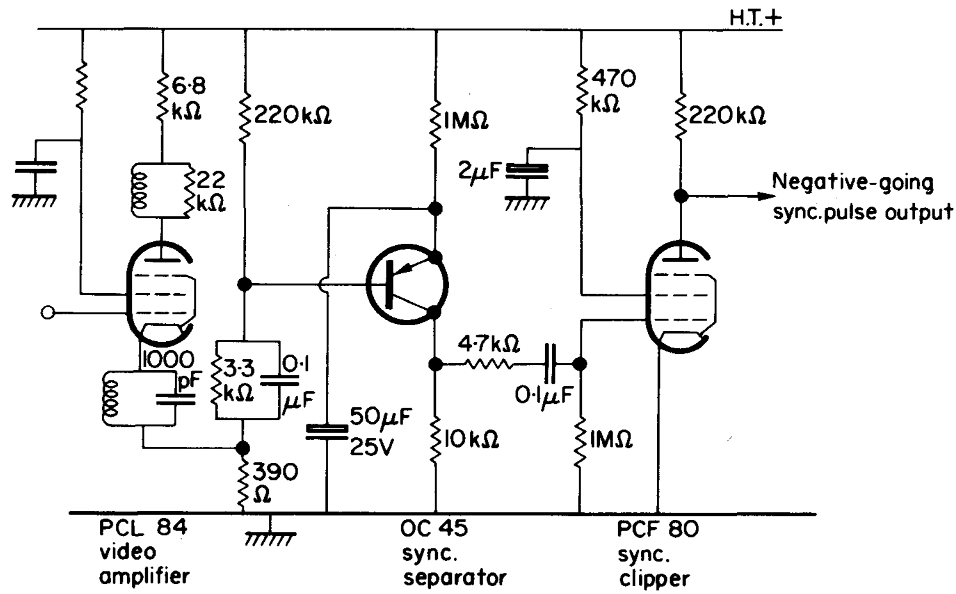



Fig. 10.7 Transistor sync. separator in use in a hybrid receiver

The first point to note is that the transistor is drawn 'upside-down'. This is necessary because the collector has to be negative to emitter and hence the collector load returns to chassis and the emitter resistor to H.T.

The required positive-going video signal is obtained from the video amplifier cathode. As far as the sync. separator is concerned, the video amplifier works as a cathode follower; i.e. the video amplifier receives a positive-going video signal input from the vision detector and develops a signal of similar amplitude and like polarity at its cathode. An immediate advantage here is that the sync. separator circuit is entirely divorced from the anode video signal path to the c.r.t.

The bias to the transistor base is seen to be a combination of a fixed bias from a potential divider and auto-bias from a C.R. network in the signal input connection.

Positive-going voltage pulses at the collector are clipped by application to a short grid base pentode clipper, which produces negative-going voltage pulses at its anode.

# 11

# Differentiators and Integrators

It is the difference in time duration of the line and field sync. pulses which makes it possible to sort them. In effect this difference of time duration is converted to a difference in final pulse amplitude. It is then easy to bias a valve or transistor so that it passes the larger amplitude pulse but rejects the smaller.

Circuitry for sorting the pulses makes use of differentiating and integrating networks. In their simplest form these are merely circuits consisting of one capacitor and one resistor. The input is applied across the two components connected in series. If the output is taken from the resistor the circuit is a *differentiator*; if taken from the capacitor it is an *integrator*. The reason for so calling them stems from a consideration of what factors direct the shape of the output waveform in each case. In the former arrangement, if the time constant of the R.C. network is short compared with the time duration of the applied pulses, then the output across the resistor resembles that which would be arrived at by subjecting the input waveform to the mathematical process of differentiation. In the latter case, if the time constant of the network is long compared with the pulse durations, then the output approximates to that arrived at by the process of integration of the applied waveform. Readers with a knowledge of the calculus may wish to pursue this aspect further, but such a knowledge is not required for understanding what follows.

It is very necessary, however, that a technician handling electronic circuitry which is processing pulses should be familiar with the behaviour of these basic networks. Towards this end careful attention is paid in this chapter to the effect of such circuits upon the television sync. pulse waveform. A simplified step-by-step approach is used to build up a realistic picture of the waveforms produced. The aim is simply to demonstrate that the circuits do what they are reputed to do, and to show a method of predicting what output waveform may be expected from a given network when fed with a pulse waveform. Having grasped the principles of these basic circuit elements, the remainder of the processing circuitry is straightforward.

Fig. 11.1 shows a differentiator and an integrator connected between anode and cathode of a pentode sync. separator. The differentiator is shown with a time-constant of about one-tenth of the duration of a line pulse, and the integrator has a time constant equal to the duration of one field pulse.

Whilst these are satisfactory values there is nothing critical about them and much variation will be found. It will be noticed that the time constants are often such as to disqualify the network from being accurately described as a differentiator or an integrator. A differentiator only truly 'differentiates' when its time-constant is short compared with the pulse duration, and an integrator only truly 'integrates' when its time-constant is long compared with the pulse time. This is, however, a rather pedantic point and in practice all such circuits are so described, despite their time constants, as a convenient means of identifying and grouping

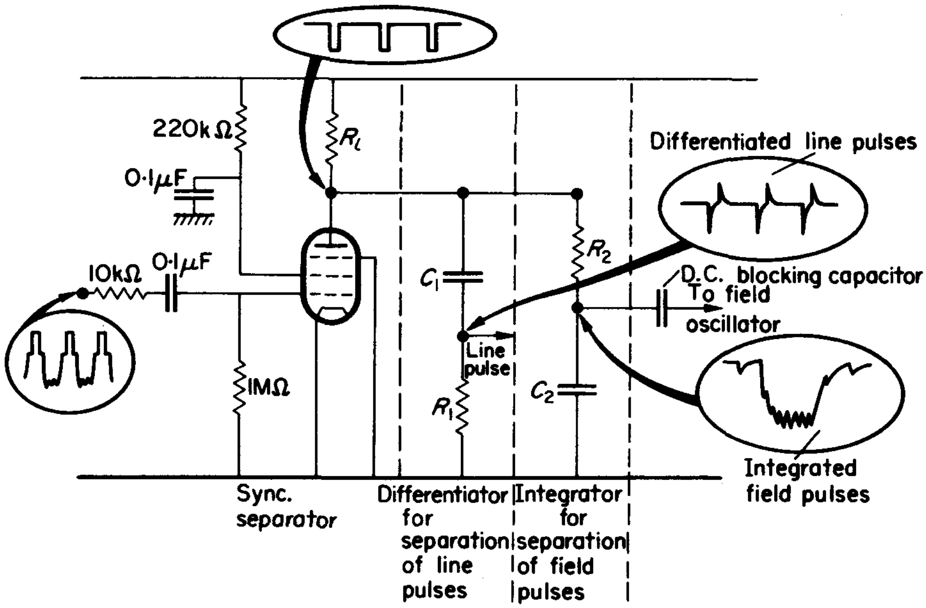



Fig. 11.1 Showing a sync. separator followed by basic differentiating and integrating networks to separate out the line and field pulses

*Note* typical time-constants:

(i) Line-pulses differentiator for British 405 signal (line pulse duration = 10 μs)

$$C_1 = 10\text{ pF} \qquad R_1 = 100\text{ k}\Omega$$

$$\begin{aligned}\text{Time-constant} &= C_1R_1 \text{ seconds}\\ &= 10\times10^{-12}\times100\times10^3\times10^6\ \mu\text{s}\\ &= \underline{1\ \mu\text{s}}\end{aligned}$$

(ii) Line-pulse differentiator for British 625 signal (line pulse duration = 4·7 μs)

$$C_1 = 5\text{ pF} \qquad R_1 = 100\text{ k}\Omega$$

$$\begin{aligned}\text{Time-constant} &= C_1R_1 \text{ seconds}\\ &= 5\times10^{-12}\times100\times10^3\times10^6\ \mu\text{s}\\ &= \underline{0{\cdot}5\ \mu\text{s}}\end{aligned}$$

(iii) Field-pulse integrator for 405 signal (field pulses = 40 μs duration)

$$C_2 = 200\text{ pF} \qquad R_2 = 200\text{ k}\Omega$$

$$\begin{aligned}\text{Time-constant} &= C_2R_2 \text{ seconds}\\ &= 200\times10^{-12}\times200\times10^3\times10^6\ \mu\text{s}\\ &= \underline{40\ \mu\text{s}}\end{aligned}$$

(iv) Field-pulse integrator for 625 signal (field pulses = 27·3 μs duration)

$$C_2 = 200\text{ pF} \qquad R_2 = 150\text{ k}\Omega$$

$$\begin{aligned}\text{Time-constant} &= C_2R_2 \text{ seconds}\\ &= 200\times10^{-12}\times150\times10^3\times10^6\ \mu\text{s}\\ &= \underline{30\ \mu\text{s}}\end{aligned}$$

them. Sometimes a term such as 'partial differentiator' is used to indicate that the time-constant is longer than, or comparable to, the pulse time, so that the resulting waveform is only partially differentiated.

The action of these circuits will now be investigated carefully.

**The line pulse differentiator**

The sync. separator anode rests at H.T. potential between pulses when the valve is non-conductive, and drops to some lower potential for the duration of the pulses. As a first approximation therefore we may represent the situation facing a network connected between anode and chassis, by the simplified diagram of Fig. 11.2(b). Here the network is shown switched between two different voltage levels. For convenience these have arbitrarily been taken as 180 V and 100 V. The resistance of the anode load and of the valve itself have been ignored for this purpose. These factors must naturally affect the output waveform, but the result obtained from this simplified approach bears a close resemblance to the actual one met in practice and the method serves as a useful vehicle for gaining the necessary understanding.

The first point to notice is that the polarity of the output voltage across $R$ is positive with respect to chassis when the capacitor is charging from the lower towards the higher level, but negative to chassis when it is discharging from the higher to the lower potential. A useful device to remember is that when electrons are passing through a resistor, the end they enter by is driven negative to the end they leave by. When $C$ is charging, electrons have to leave the 'top' plate and pile up on the 'bottom' plate. Their direction is thus up through $R$, and the top of $R$ is driven positive to chassis. When $C$ discharges, electrons travel from its 'bottom' plate towards the 'top' plate; i.e. downwards through $R$ making the top negative to chassis.

Secondly, note that at any time the instantaneous voltage across $R$ must be the difference between the applied voltage and the voltage to which the capacitor is charged, i.e. $V_r = (V_b - V_c)$ volts.

Suppose that the switch has rested for some time at position 'b'. $C$ will be charged to 100 V and no current will be flowing through $R$, i.e. $V_r = 0$.

If the switch is moved to position 'a', $C$ now begins to charge from 100 V towards 180 V. At the instant the switch is thrown, $V_r = (180 - 100)$ V $= +80$ V. However $C$ charges according to the normal exponential law and the voltage gap between $V_b$ and $V_c$ closes. Thus, as $V_c$ rises the initial $+80$ V across $R$ dies away exponentially. Fig. 11.2(c) shows the way in which $V_c$ and $V_r$ vary.

In Fig. 11.2(d) we assume that $S$ has rested for some while at 'a' and that $C$ is charged to

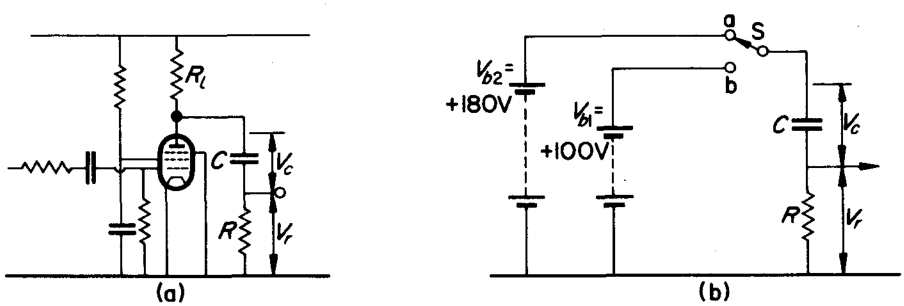



Fig. 11.2(a) Sync. separator followed by line-pulse differentiator

Fig. 11.2(b) Simplified circuit for explanatory purposes (see text)

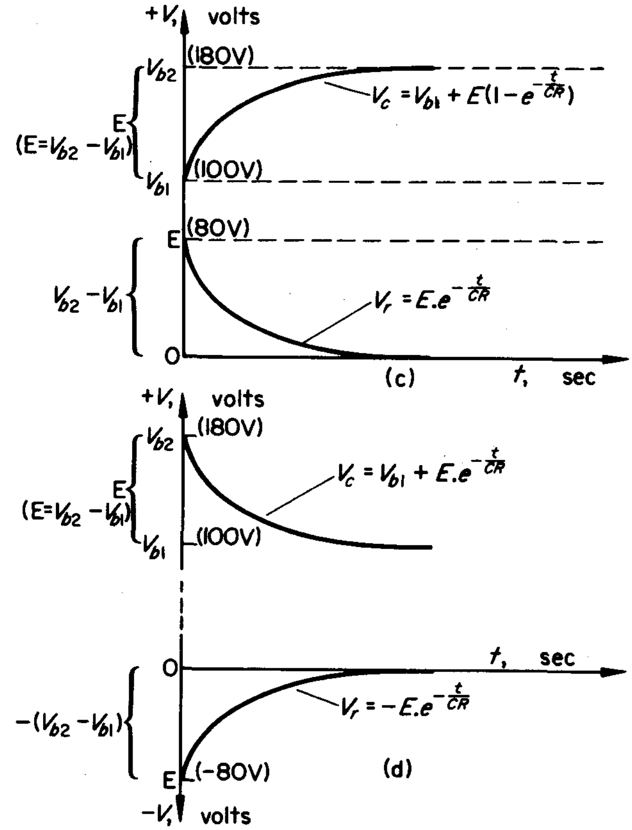



Fig. 11.2(c) When $C$ is charging to a higher voltage, electrons flow up through $R$ and $V_r$ is positive to chassis. The instantaneous values of $V_c$ and $V_r$ are given by the formulae shown.

Fig. 11.2(d) When $C$ is discharging to a lower voltage level, electrons flow down through $R$ making $V_r$ negative to chassis. Again, the formulae governing the instantaneous values of $V_c$ and $V_r$ are shown.

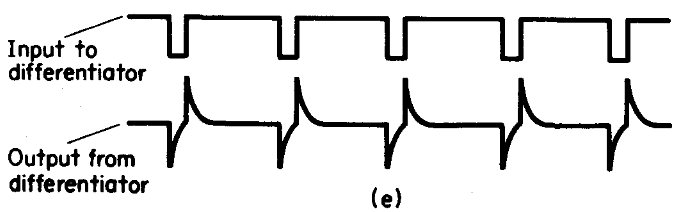



Fig. 11.2(e) Showing the output waveform produced by a differentiator during normal picture lines when fed with the idealised output from a sync. separator

180 V. Again $V_r = 0$ V. When $S$ is moved to position 'b' the instantaneous voltage across $R$ is given by $V_r = (100 - 180)$ V $= -80$ V. $C$ now discharges from 180 V towards 100 V and the negative voltage across $R$ falls away as the gap between $V_b$ and $V_c$ narrows.

In Fig. 11.2(e) the idealised input to the differentiator during normal picture lines is shown. Below this waveform the typical output from a line-pulse differentiator is drawn.

The way in which this waveform comes about is established in Fig. 11.3. For convenience the British 405-line waveform is considered. Here the line pulses are of approximately 10 μs duration and the inter-pulse periods 90 μs. The differentiator has a time-constant of 1 μs, and the simple switching circuit of Fig. 11.2(b) is again used as a basis for discussion.

At time $t_1$ towards the end of the picture line, we assume the capacitor to be fully charged to 180 V, and $V_r=(V_b-V_c)=(180-180)$ V$=0$ V.

At $t_2$ a line sync. pulse arrives and the applied voltage drops sharply to 100 V. Thus $V_r=(V_b-V_c)=(100-180)$ V$=-80$ V.

Between $t_2$ and $t_5$, $C$ discharges from 180 V towards 100 V. In a time equal to the time-constant time of $CR$ seconds$=1$ μs, the voltage gap of 80 V facing $C$ will have diminished to 37% of 80 V, i.e. $C$ will have lost 63% of its excess charge. At this moment, shown as $t_3$ in the

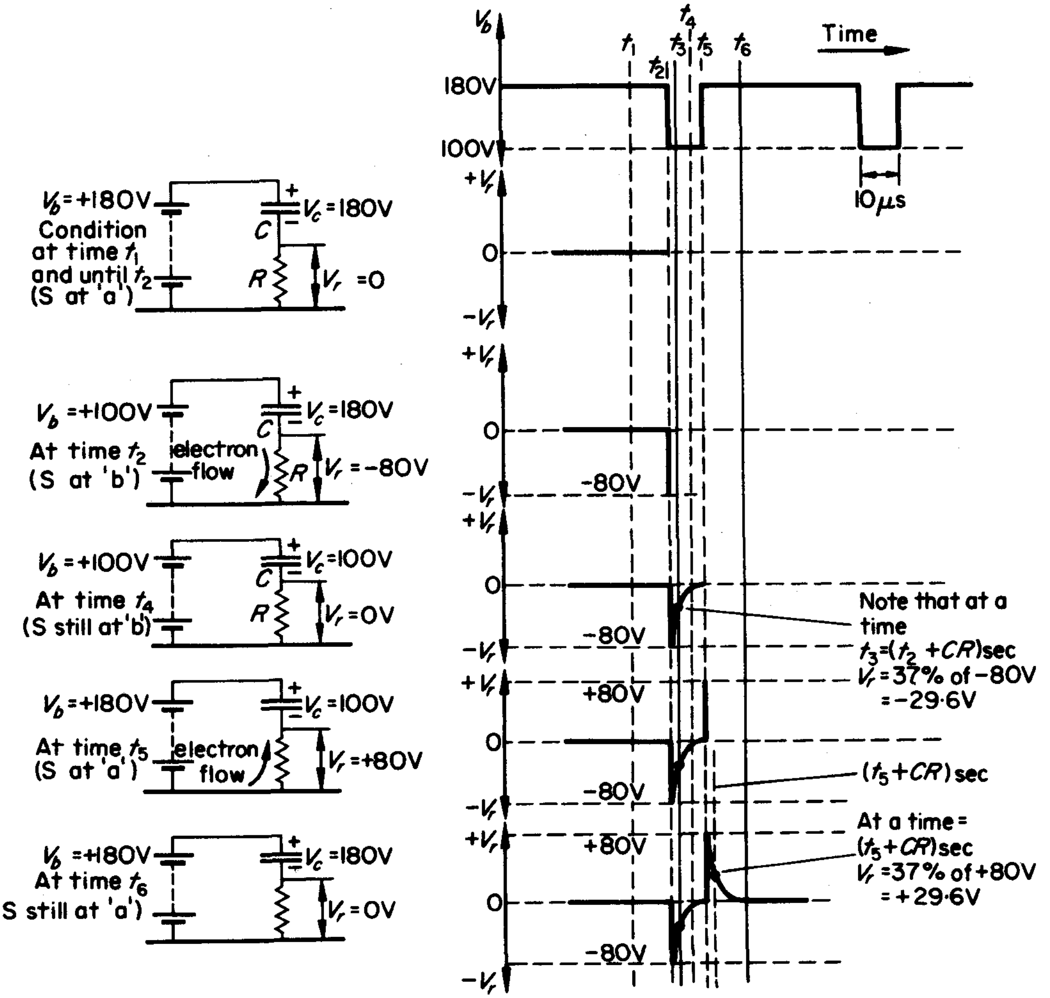



Fig. 11.3 Illustrating how a differentiating circuit produces its characteristic output from a square pulse input
The time-constant of the differentiator is assumed to be approximately one-tenth of the time duration of the sync. pulse.

diagram, $V_r=(V_b-V_c)=(100-129{\cdot}6)=-29{\cdot}6$ V. Just before the end of the line pulse, at a time $t_4$, $V_c$ will equal $V_b$ and $V_r=0$ V. At $t_5$ the pulse terminates and the applied voltage jumps up to 180 V. This takes the voltage across $R$ to $V_r=(V_b-V_c)=(180-100)=+80$ V.

$C$ immediately starts to charge from 100 towards 180 V and at a time $(t_5+1\ \mu s)$ it will have closed 63% of the 80 V gap, i.e. at this time $V_r=(V_b-V_c)=(180-150{\cdot}4)$ V$=+29{\cdot}6$ V. At $t_6$, long before the next line pulse, $V_c$ will rest virtually at 180 V and $V_r$ will be for all practical purposes zero.

The diagrams of Fig. 11.3 trace these events step by step. The differentiated waveform is seen to have sharp spikes on the leading and trailing edges of the line pulses. The sharp negative-going pulse on the leading edge is used to trigger the line timebase.

The effects of the valve, the anode load and of shunt capacitance do modify the practical waveform to some extent, but something very like this theoretical one does in fact appear, the chief difference being that the positive-going trailing edge spike is of lower amplitude and occupies a greater width. It should be noticed that the effective time-constant is different on the charge and discharge legs in the practical circuit. When the sync. separator is cut off, $C$ charges to the H.T. potential via the anode load which gives a time-constant of $C(R_l+R)$ seconds. When the valve conducts (on the arrival of a pulse), $C$ may discharge through the low conductive resistance of the valve, and here the time-constant is more nearly that due to $C$ and $R$ alone. In addition to this, as was shown in Fig. 9.4(d), the presence of shunt capacitance across the anode circuit causes a pronounced curvature on the trailing edge of the anode pulse. The *input* to the differentiator is therefore not a perfect squared pulse.

In practice the combined effect of these factors is that the leading-edge spike is sharper and of greater amplitude than the spike on the trailing edge.

Later on in this chapter, a partial differentiator for a field pulse application is studied. Here the time-constant is not short compared with the pulse time, so that, having been shifted sharply from one level to another on the pulse leading edge, $V_r$ has not diminished to the start (zero) line by the end of the pulse. This substantially alters the effect of the circuit, as will be seen.

Fig. 11.4 shows how the line-pulse differentiator responds to the video waveform during a field sync. pulse chain. It is the leading edge of a normal differentiated line pulse which is used for triggering the line timebase. A study of the diagram shows that the leading edge of field

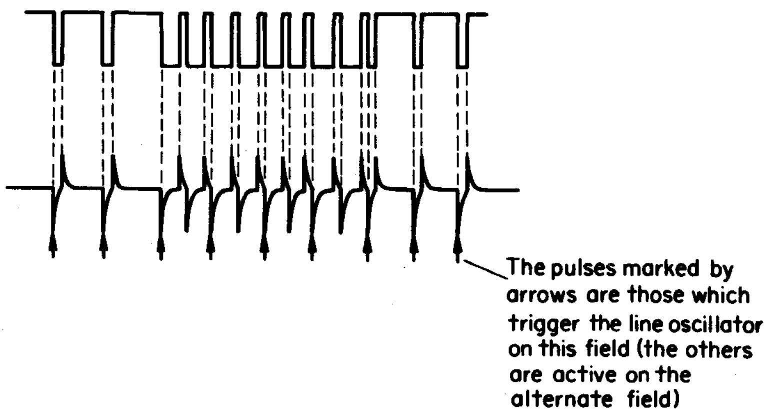



Fig. 11.4 Showing the waveform produced by a line-pulse differentiator during the field-pulse sequence of the British 405-line signal

pulses produces the necessary negative-going pulses to keep the line timebase synchronised during the field pulse sequence. During this sequence, because of the nature of the field pulse waveforms, line pulses appear at twice their normal rate.

Care has to be taken in line oscillator design to ensure that only alternate pulses trigger the timebase. It is possible for a condition to arise such that the oscillator is triggered by both the half-line and the end of line pulses, so that it runs at twice its normal speed for the duration of the field pulse chain. The upset caused by this can persist for several lines after the end of the field pulse chain. This causes a wobbling disturbance at the top of the raster, with shorter than normal lines tending to pull out to one side. The design requirement is that the line sync. pulses, whilst being of sufficient amplitude to do their job effectively, are nevertheless too small to trigger the line oscillator when it is in the middle of its scanning stroke. Even when this is so, however, receivers employing 'direct' as distinct from 'flywheel' synchronisation, often show a disturbance at the top of the raster. This is due to the difference in the nature of the sync. pulse input waveform to the line oscillator during the field pulse sequence, causing changes in the voltage levels on time constant capacitors in the oscillator circuitry. The result of this is a change in the amplitude and position of the uppermost lines on the raster.

**The field pulse integrator**

This circuit has a long time-constant, with the result that it pays less attention to line pulses which are short than to field pulses which are considerably longer. Field pulses are approximately six times as long as line pulses in the British 625-line signal and four times as long in the 405-line signal.

The circuit contrives to add up (i.e. to integrate) the successive field pulses, so producing a large single pulse of duration equal to the total length of the field pulse chain. Fig. 11.6 and Fig. 11.7 show the forms which the derived integrated waveforms take on the field pulse trains of the 405-line and 625-line signals respectively. In order to demonstrate how these wave shapes come about, the same technique is used as for the differentiator. The circuit of Fig. 11.5(a) is replaced by the simplified circuit of Fig. 11.5(b). It is now merely a question of working

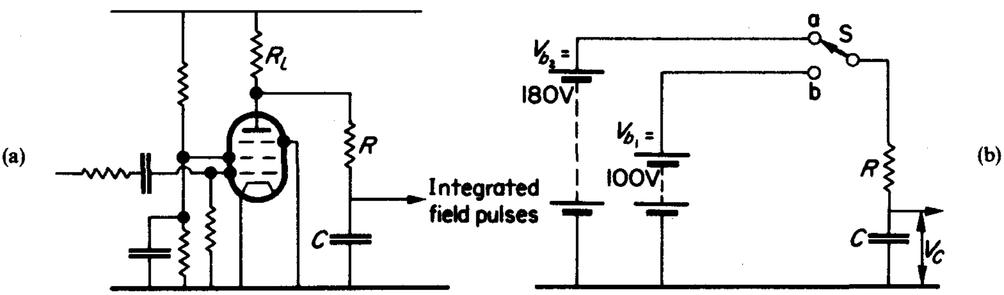



Fig. 11.5(a) Sync. separator followed by field-pulse integrator

Fig. 11.5(b) Simplified circuit used to build the waveforms of Fig. 11.6 and Fig. 11.7. The time-constant (*CR* seconds) is taken as 30 μs for the 625-line signal and 40 μs for the 405-line signal.

out what happens as the switch moves to and fro between the two voltage levels. As before, the potential at the sync. separator anode is taken as 180 V between pulses, dropping to 100 V during the pulses.

The waveform produced on odd and even fields is deduced for both the 405-line and 625-line signals. The result brings out a very important difference between these two signals. In order to spotlight this difference the 405-line signal is dealt with first.

*The integrated 405-line field-pulse train* (Fig. 11.6)

A little patience is needed in building this diagram, but the exercise is most worthwhile. Subsequently, if the reader is in a position to do so, the result may be verified experimentally and the waveform seen on an oscilloscope. This may be done by strapping an integrator between anode and chassis of the sync. separator in a T.V. receiver which is tuned to a signal, and connecting the oscilloscope across the capacitor. The leads to the line and field oscillators should be disconnected from the sync. separator to prevent fed-back oscillator waveforms from upsetting the trace.

The formulae shown on Fig. 11.2 govern the charge and discharge of the capacitor. The situation is identical except that here it is the capacitor voltage which is of interest and not the waveform across the resistor.

As a starting point the factor $e^{-\frac{t}{CR}}$ should be worked out for the time intervals met along the waveform. These are 90 μs, 10 μs, 40 μs and one dimension of 60 μs. The results are tabulated under the diagram. For each successive calculation the 'voltage gap' facing $C$ (i.e. the difference between the battery and capacitor voltages) when the switch moves over at the start of an interval, is multiplied by the factor appropriate to the length of that interval. This step leads to the corresponding voltage gap at the end of the interval.

The net voltage $V_c$ at the end of the interval is then found by adding the final gap voltage to 100 V for downward movements of $V_c$ (i.e. when the switch is at $b$), or subtracting the final gap voltage from 180 V on upward movements of $V_c$ (i.e. when the switch is at $a$).*

*Pulse train following end of even fields*

<u>At $t_1$</u> $S$ is at 'a' and has rested there for some 80 μs or so. It may be assumed as a first approximation that $C$ is fully charged to the applied 180 V. Thus $V_c = +180$ V.

<u>At $t_2$</u> $S$ moves to 'b', i.e. $V_b$ drops to 100 V on the arrival of a line sync. pulse. $C$ begins to discharge from 180 V down towards 100 V. In the 10 μs interval between $t_2$ and $t_3$ the 80 V net voltage across $C$ will reduce to $0{\cdot}8 \times 80 = 64$ V. This means that the actual voltage across $C$ reduces to 164 V in this time.

* This may be expressed as generalised formulae as follows:—

If $V_c'$ is the capacitor voltage at the start of an interval of $t$ seconds
$V_c$ the corresponding voltage at the end of the interval
$V_g$ is the voltage gap at the start of the interval
$V_{b2}$ and $V_{b1}$ are the higher and lower applied voltages respectively, then

On downward movements of $V_c$ when $V_{b1}$ is the applied voltage (i.e. when the switch is at $b$):—

$$\underline{V_c = V_{b1} + V_g . e^{-\frac{t}{CR}}}$$

On upward movements of $V_c$ when $V_{b2}$ is the applied voltage (i.e. when the switch is at $a$):—

$$V_c = V_c' + V_g(1 - e^{-\frac{t}{CR}})$$

$$\text{but } V_c' = V_{b2} - V_g$$

$$\therefore \underline{V_c = V_{b2} - V_g . e^{-\frac{t}{CR}}}$$

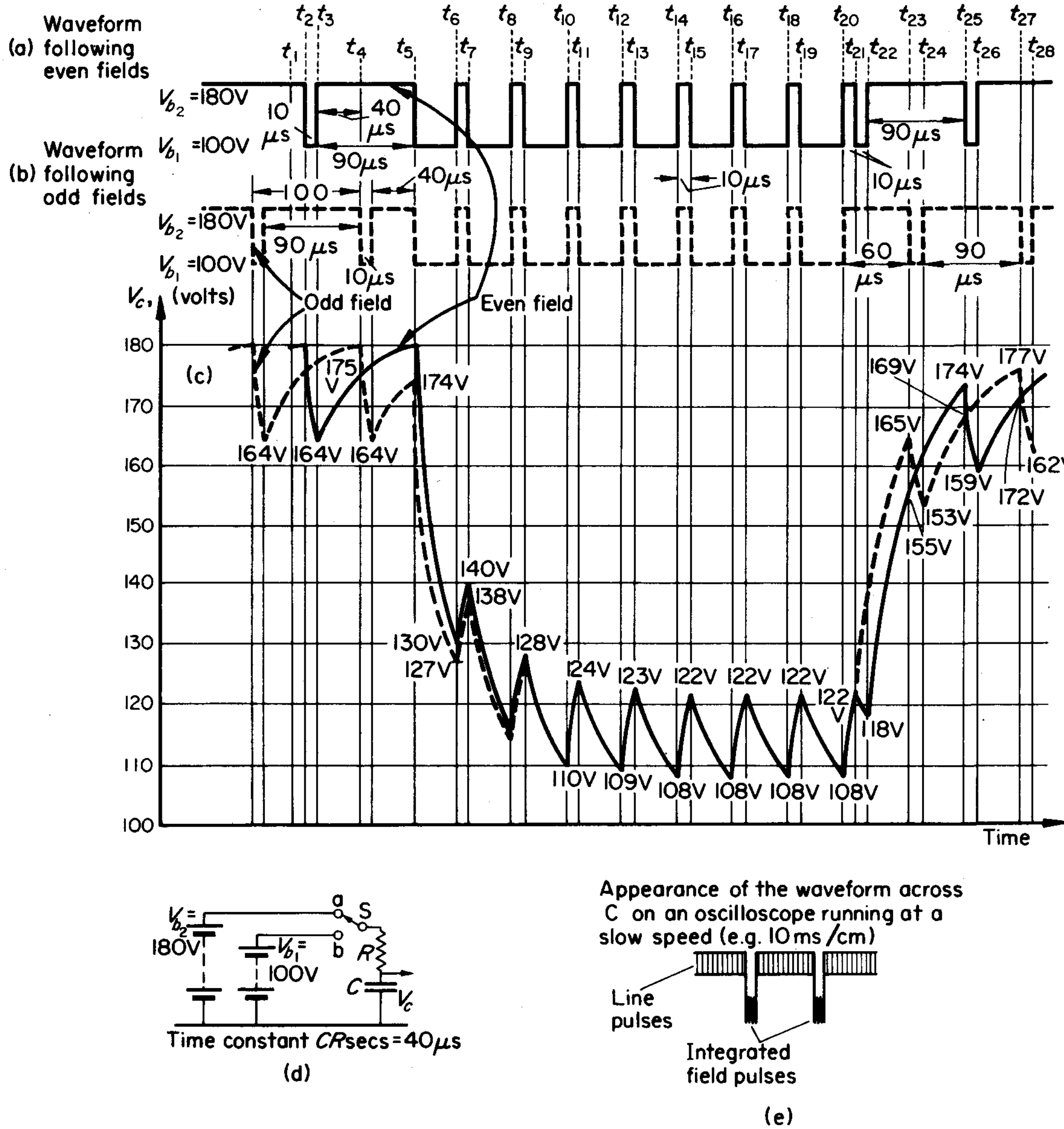



Fig. 11.6 Illustrating how an integrating circuit produces its characteristic output from the pulse trains which follow even and odd fields. (For line numbers, if required, see Fig. 3.4, page 34.)

Table of values of the factor $e^{-\frac{t}{CR}}$ where $t$ is the time interval and $CR = 40\ \mu s$

| | |
|---|---|
| When $t = 10\ \mu s$ | $e^{-\frac{t}{CR}} = e^{-\frac{10}{40}} = e^{-0 \cdot 25} = 0 \cdot 7788 \simeq 0 \cdot 8$ |
| When $t = 40\ \mu s$ | $e^{-\frac{t}{CR}} = e^{-\frac{40}{40}} = e^{-1} = 0 \cdot 3679 \simeq 0 \cdot 37$ |
| When $t = 60\ \mu s$ | $e^{-\frac{t}{CR}} = e^{-\frac{60}{40}} = e^{-1 \cdot 5} = 0 \cdot 2231 \simeq 0 \cdot 2$ |
| When $t = 90\ \mu s$ | $e^{-\frac{t}{CR}} = e^{-\frac{90}{40}} = e^{-2 \cdot 25} = 0 \cdot 1055 \simeq 0 \cdot 1$ |

At $t_3$ — $S$ moves to 'a'. $V_b$ returns to 180 V. $C$ begins to charge from 164 V towards 180 V. The interval between $t_3$ and $t_5$ is 90 μs. In the first 40 μs of this interval, by time $t_4$, the $(180-164)$ V$=16$ V net voltage gap facing $C$ will narrow by 63% of 16V$=11$ V. Thus $V_c=(164+11)$ V$=175$ V at this time. By $t_5$, $V_c$ again substantially equals the applied 180 V. (There is a small error in assuming this, but it makes no significant difference to the shape of the final pulse drawn.)

At $t_5$ — $V_b$ drops to 100, and stays there for the 40 μs duration of the first field pulse. In this time $C$ discharges from 180 V down towards 100 V and the 80 V voltage gap is closed to 37% of 80 V$=30$ V by time $t_6$. Hence, by $t_6$, $V_c$ has fallen to 130 V.

At $t_6$ — $V_b$ goes up to 180 V. The voltage gap facing $C$ is now $(180-130)=50$ V. In the 10 μs between $t_6$ and $t_7$ $C$ charges from 130 V towards 180 V and the 50 V gap narrows to $0{\cdot}8\times50$ V$=40$ V. Thus at time $t_7$, $V_c$ has reached $(180-40)$ V$=140$ V.

At $t_7$ — $V_b$ drops to 100 V, and stays there for 40 μs. In this time $C$ discharges from 140 V down towards 100 V. By time $t_8$ the 40 V gap faced by $C$ has reduced to 37% of 40 V$=15$ V. At time $t_8$, $V_c$ rests at 115 V.

At $t_8$ — $V_b$ rises to 180 V. This faces $C$ with a gap of $(180-115)$ V$=65$ V. In the 10 μs between $t_8$ and $t_9$ this gap closes to 0·8 of 65 V$=52$ V. Thus at $t_9$ $V_c=(180-52)$ V$=128$ V.

At $t_9$ — $V_b$ drops to 100 V and stays there for 40 μs. In this time $C$ discharges from 128 V towards 100 V. By $t_{10}$ the 28 V gap has reduced to 37% of 28 V$=10$ V. Thus at $t_{10}$, $V_c=110$ V.

At $t_{10}$ — $V_b$ rises to 180 V, and stays there for 10 μs. During this time $C$ charges from 110 V up towards 180 V. By time $t_{11}$ the 70 V gap has closed to $0{\cdot}8\times70$ V$=56$ V, which means that $V_c=(180-56)$ V$=124$ V.

At $t_{11}$ — $V_b$ drops to 100 V and stays there for 40 μs. In this time $C$ discharges from 124 V down towards 100 V. By $t_{12}$ the 24 V gap has reduced to 37% of 24 V$=9$ V. Thus at $t_{12}$ $V_c=109$ V.

At $t_{12}$ — $V_b$ rises to 180 V. $C$ charges from 109 V towards 180 V and the 71 V gap reduces to $(0{\cdot}8\times71)$ V$=57$ V in the 10 μs between $t_{12}$ and $t_{13}$. Thus at $t_{13}$, $V_c=(180-57)$ V$=123$ V.

At $t_{13}$ — $V_b$ drops to 100 V and stays there for 40 μs. In this time $C$ discharges from 123 V down towards 100 V. By $t_{14}$ the 23 V gap has reduced to 37% of 23 V$=8$ V. Thus at $t_{14}$, $V_c=108$ V.

It is now seen that the circuit has reached a state of equilibrium, such that the negative-going movements during the field pulses are balanced by equal positive-going movements during the half-line pulses. This is quite logical since the voltage gap facing $C$ on upward excursions is much greater than the gap facing it during the downward ones. For this reason the rate of voltage change during a 10 μs positive-going movement is faster than the change during the longer 40 μs negative-going movements, and the actual voltage changes are equal.

The integrated field pulse is thus characterized by a saw-edged bottom. This is clearly seen on an oscilloscope. When the 'X-gain' is turned down so that the pulse 'closes up' the waveform seen takes on the form shown in Fig. 11.6(e). This appears as a rectangular pulse with a filled-in tip. Seen like this the pulse looks better than it really is. The leading edge appears as a straight drop, when, as Fig. 11.6(c) shows, it is in reality a somewhat sloping edge bearing serrations corresponding to the half-line pulses.

The trailing edge of the waveform of Fig. 11.6(c) is deduced in the same way and the reader is recommended to verify the points plotted.

The corresponding waveform for odd fields is also plotted on the same diagram. It is

immediately apparent that the integrated odd and even field pulses are not identical. Differences occur both on the leading and the trailing edges. Both of these differences cause interlacing difficulties.

Taking the leading edge discrepancy first, the difference is due to the influence of the last line sync. pulse which occurs before the start of the field pulse sequence. After even fields there is an interval of 90 μs between the last line pulse and the first field pulse. During this time the negative going voltage change across $C$ due to the line pulse has been almost entirely restored and the first field pulse finds $V_c$ at +180 V.

After odd fields however, the first field pulse occurs in the middle of a line; separated by only 40 μs from the last line pulse. In this case the voltage change caused by the last line pulse has been only partially restored when the first field pulse arrives.

For this reason the downward excursion of the voltage $V_c$ on the first field pulse, starts from different initial levels after odd and even fields, and hence the leading edges of the integrated field pulses cannot coincide. As the pulse chain proceeds, the disparity diminishes, until by the end of the second field pulse the two integrated waveforms are virtually coincident.

The trailing edge suffers in a similar way. On both odd and even fields the voltage across $C$ rests at the same level at the end of the last field pulse. Thereafter the waveforms again separate because the voltage changes presented to the integrator differ considerably. This is self-explanatory from the diagram.

Before discussing why it is that both of these discrepancies can cause the quality of interlace to be impaired, it is instructive to study the corresponding waveform for the 625-line signal.

*The integrated 625-line field pulse chain* (Fig. 11.7)

The purpose and effectiveness of the equalising pulses is now apparent. The integrated waveforms shown were built by the same step-by-step method demonstrated for the 405-line signal. The student is strongly recommended to take the trouble to satisfy himself by approximate calculation that he sees clearly how the waveforms are built up.

A study of the 625-line signal shows that the inclusion of equalising pulses before and after the field pulses insulates the integrator from the differences which must occur as a result of inserting the field pulse waveform in the middle of one line on one field, but at the end of a line on alternate fields.

On odd fields the first equalising pulse appears in the middle of line 623. The interval between the leading edge of this pulse and the leading edge of the last line sync. pulse to occur before it, is 31·7 μs. The integrated waveform of Fig. 11.7 shows that the capacitor has not recovered from the effect of the last line pulse, when this first equalising pulse appears.

On even fields the first equalising pulse occurs at the end of line 310 and a full line period of 64 μs has elapsed since the last line pulse. In this case the capacitor has virtually fully recovered from the effect of the last line pulse when this first equalising pulse arrives.

There is thus a difference in the integrated waveforms at the onset of the equalising pulses. Thereafter, however, the applied signal waveforms are identical, and by the end of the two and a half line equalising pulse period, the effect on the integrator of the initial discrepancy has disappeared, and the waveforms across $C$ are coincident. The integrator now operates on the broad field pulses to produce exactly the same integrated field pulse waveform on both fields. This coincidence of the waveforms persists up to the end of the post-field pulse equalising pulses.

There is a further difference at this point because the trailing edge of the last equalising pulse is separated from the leading edge of the next succeeding line pulse by 30 μs on the waveform

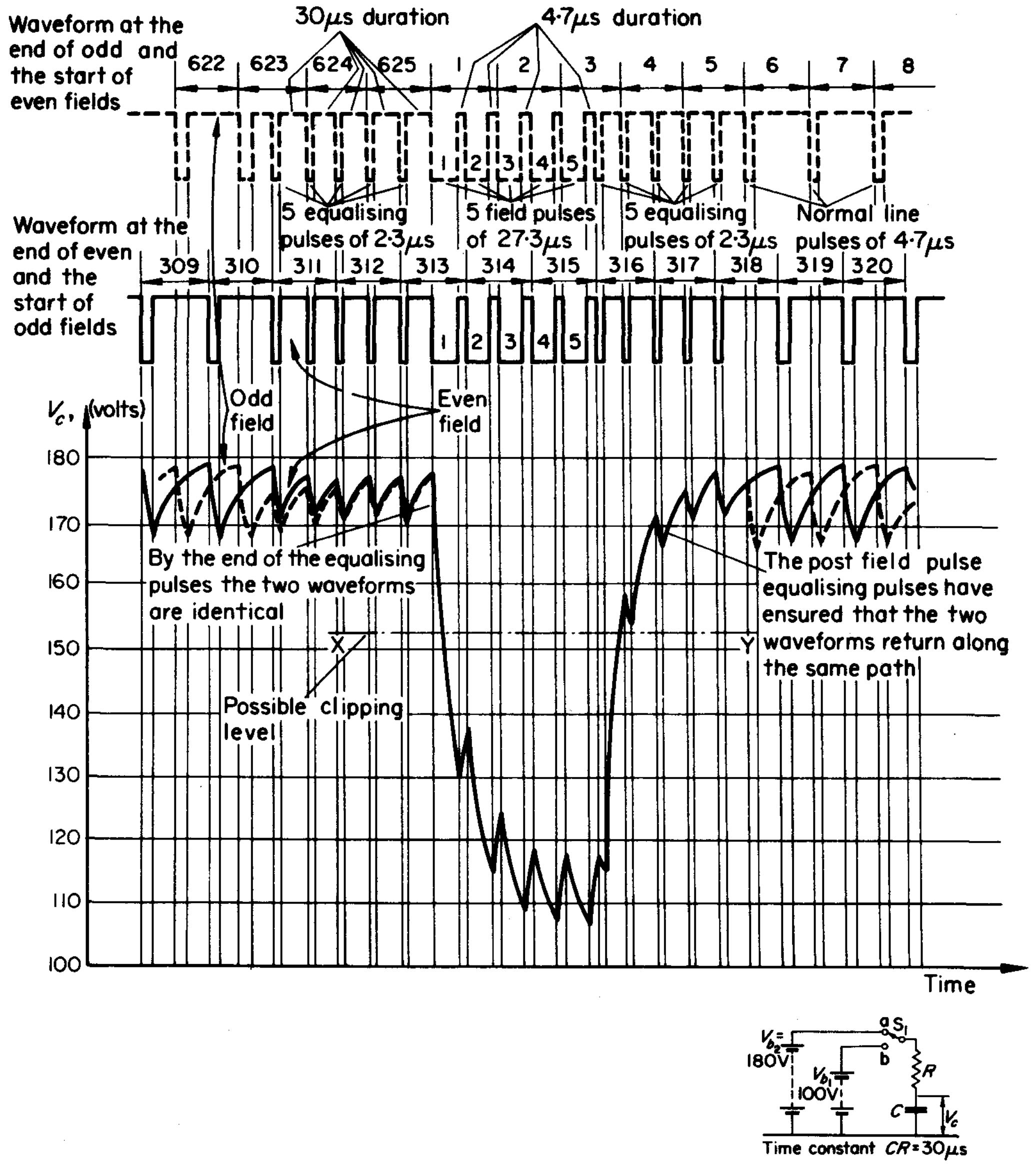



Fig. 11.7 Illustrating how an integrating circuit produces its characteristic output from even and odd field-pulse trains on a 625-line signal

Table of values of the factor $e^{-\frac{t}{CR}}$ where $CR = 30$ µs

| | |
|---|---|
| When $t = 2{\cdot}3$ µs | $e^{-\frac{t}{CR}} = e^{-\frac{2\cdot3}{30}} = e^{-0\cdot077} \simeq e^{-0\cdot08} \simeq 0{\cdot}923$ |
| When $t = 4{\cdot}7$ µs | $e^{-\frac{t}{CR}} = e^{-\frac{4\cdot7}{30}} = e^{-0\cdot157} \simeq e^{-0\cdot16} \simeq 0{\cdot}85$ |
| When $t = 30$ µs | $e^{-\frac{t}{CR}} = e^{-\frac{30}{30}} = e^{-1} = 0{\cdot}3678 \simeq 0{\cdot}37$ |
| When $t = 60$ µs | $e^{-\frac{t}{CR}} = e^{-\frac{60}{30}} = e^{-2} = 0{\cdot}13534 \simeq 0{\cdot}135$ |

which follows the end of odd fields, but by 61·7 μs on the one which follows the end of even fields.

The differences which occur at the start of the equalising pulses which precede the field pulses, and at the end of those which follow the field pulses are of no consequence at all. This is because the integrated field pulse waveform is normally clipped off, for example along the line XY on Fig. 11.7, and the voltage changes across $C$ due to line pulses are not seen by the field oscillator circuit at all.

**The field pulse differentiator**

These circuits are more properly called partial differentiators because they employ time-constants which are comparable to the time duration of the pulses being handled, rather than the shorter time constants necessary to give true differentiation. They offer an alternative method of using the transmitted information to provide synchronisation of the receiver field oscillator.

Fig. 11.8(a) shows a field pulse differentiator connected across the output of the sync. separator, and Fig. 11.8(d) shows the type of output waveform produced by such circuits.

In order to establish how the waveform comes about, the same technique is employed as was used for line pulse differentiators and field pulse integrators. Once again the circuit is

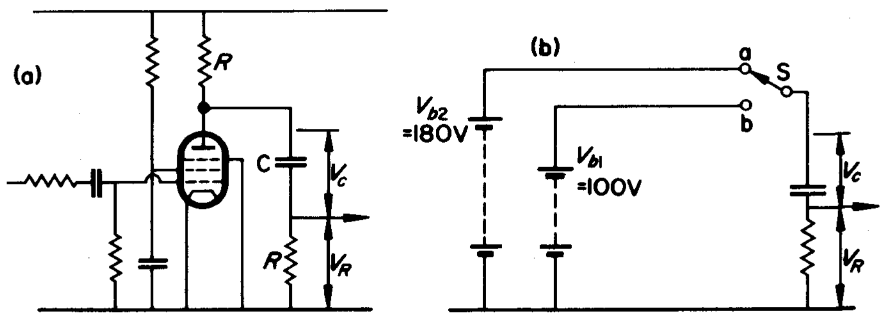



Fig. 11.8(a) Field pulse partial-differentiator circuit. The time-constant $CR$ seconds is of the same order as the time duration of one field pulse (e.g. 40 μs on the 405-line signal, or 27 μs on the 625-line signal).

Fig. 11.8(b) Simplified circuit for explanatory purposes (see text)

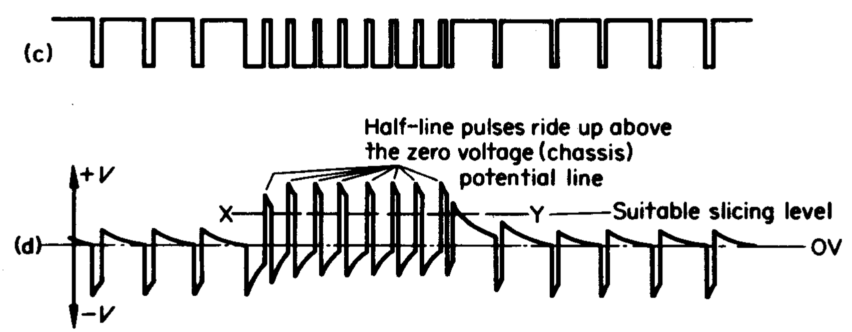



Fig. 11.8(c) Waveform at anode of sync. separator (405 signal, at the end of the even field)

Fig. 11.8(d) Waveform produced by a field pulse partial-differentiator of time constant 40 μs

simplified by imagining the network to be switched between the two arbitrarily chosen voltage levels of 180 V in the inter-pulse periods, falling to 100 V during the pulses. As before the effects of the anode load resistor, of the shunt capacitance across the anode circuit and of the resistance of the value when conductive are not taken into account. None the less, the waveform arrived at bears a close resemblance to that found in a practical working circuit.

*The partially differentiated 405-line field pulse train* (Fig. 11.9)

Suppose that at time $t_1$, just before the end of line 404 of the even field, the capacitor $C$ has fully charged to 180 V. There is then no current through $R$ and $V_r=0$. A picture of the output waveform which appears across $R$, is now built up by a step-by-step examination of the action of the circuit through the field sync. pulse sequence.

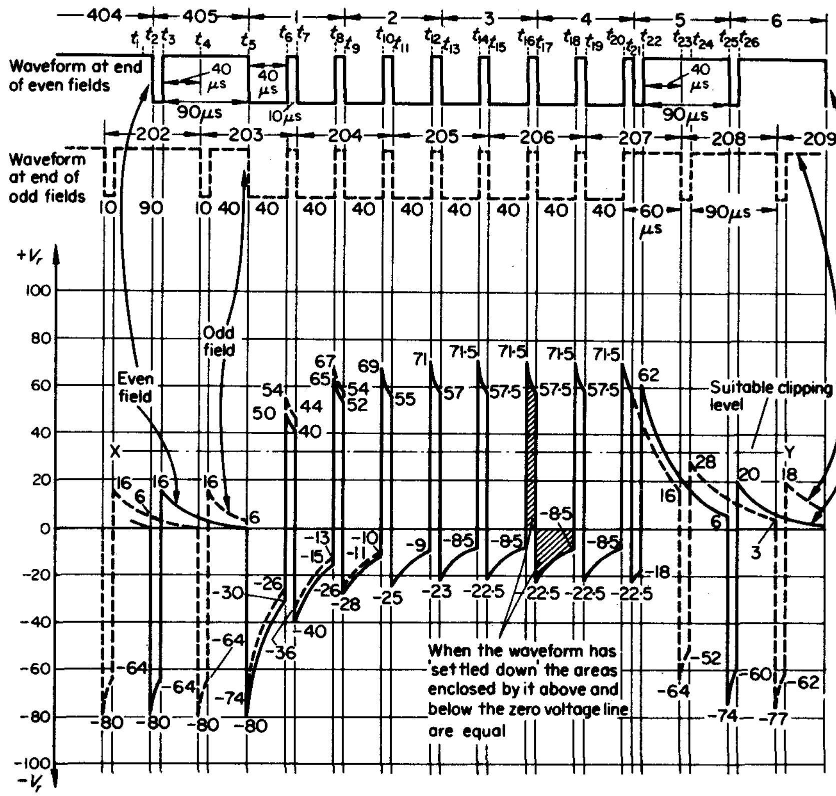
| | |
|---|---|
| At $t_2$ | $V_b$ falls to 100 V on the arrival of the line pulse. At this instant $V_r=(V_b-V_c)=(100-180)$ V $=-80$ V. |
| Between $t_2$ and $t_3$ | $V_b$ stays at 100 V for 10 µs. During this time, since the C.R. time-constant is 40 µs, $V_c$ will partially discharge to reduce the 80 V net charge upon it, to 0·8 of $80=64$ V. Thus $V_r=(V_b-V_c)=(100-164)=-64$ V. |
| At $t_3$ | $V_b$ rises to 180 V at the end of the line pulse. At this instant $V_r=(V_b-V_c)=(180-164)$ V $=+16$ V. |
| Between $t_3$ and $t_5$ | $C$ now charges to close the 16 V gap facing it. By $t_4$, i.e. after 40 µs, 63% of this gap will have been closed. This makes $V_c=(164+63\%$ of $16)$ V $=(164+10)$ V $=174$ V. Thus at $t_4$, $V_r=(V_b-V_c)=(180-174)$ V $=+6$ V. By $t_5$ it may be assumed that $V_c$ becomes fully charged to 180 V making $V_r=0$ V. |
| At $t_5$ | $V_b$ falls to 100 V. At this instant $V_r=(V_b-V_c)=(100-180)$ V $=-80$ V. |
| Between $t_5$ and $t_6$ | During this 40 µs interval, $C$ will discharge from 180 V towards 100 V. By the end of this period the 80 V gap voltage will have fallen to 37% of 80 V $=30$ V. This makes $V_c=130$ V, and $V_r=(V_b-V_c)=(100-130)$ V $=-30$ V. |
| At $t_6$ | $V_b$ rises to 180 V. At this instant $V_r=(V_b-V_c)=(180-130)$ V $=+50$ V. Thus on the leading edge of the first half-line pulse the output waveform rides up to 50 V positive to chassis, and 34 V positive to the highest potential reached during normal picture lines. |
| Between $t_6$ and $t_7$ | $C$ charges from 130 V up towards 180 V and in the 10 µs half-line pulse period the 50 V gap closes by $(0{\cdot}2\times 50)$ V $=10$ V. Thus $V_c=(130+10)$ V $=140$ V and $V_r=(V_b-V_c)=(180-140)$ V $=+40$ V. |
| At $t_7$ | $V_b$ falls to 100 V. At this instant $V_r=(V_b-V_c)=(100-140)$ V $=-40$ V. |
| Between $t_7$ and $t_8$ | $C$ discharges from 140 V down towards 100 V. In the 40 µs period the 40 V gap is reduced to 37% of 40 V $=15$ V. This makes $V_c=115$ V and thus by $t_8$, $V_r=(V_b-V_c)=(100-115)$ V $=-15$ V. |
| At $t_8$ | $V_b$ rises to 180 V. At this instant $V_r=(V_b-V_c)=(180-115)$ V $=+65$ V. Thus the leading edge of the second half-line pulse rides up to 65 V positive to chassis. |
| Between $t_8$ and $t_9$ | $C$ charges from 115 V up towards 180 V. In the 10 µs period the 65 V gap reduces by (0·2 of 65) V $=13$ V. Thus $V_r=(V_b-V_c)=(180-128)$ V $=+52$ V. |
| At $t_9$ | $V_b$ falls to 100 V. At this instant $V_r=(V_b-V_c)=(100-128)$ V $=-28$ V. |
| Between $t_9$ and $t_{10}$ | $C$ discharges from 128 V down towards 100 V. In the 40 µs period the 28 V gap reduces to 37% of 28 V $=11$ V. This makes $V_c=111$ V and $V_r=(V_b-V_c)=(100-111)$ V $=-11$ V. |

At $t_{10}$ — $V_b$ rises to 180 V. At this instant $V_r=(V_b-V_c)=(180-111)$ V $=+69$ V. Thus the leading edge of the third half-line pulse rides up to 69 V positive to chassis.

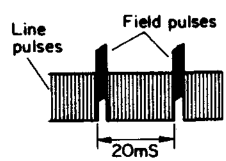
Between $t_{10}$ and $t_{11}$ — $C$ charges from 111 V towards 180 V. In the 10 μs period the 69 V gap reduces by $(0{\cdot}2\times69)$ V $=14$ V. Thus $V_c=(111+14)$ V $=125$ V and $V_r=(V_b-V_c)=(180-125)$ V $=55$ V.

At $t_{11}$ — $V_b$ falls to 100 V. At this instant $V_r=(V_b-V_c)=(100-125)$ V $=-25$ V.

Between $t_{11}$ and $t_{12}$ — $C$ discharges from 125 V down towards 100 V. In the 40 μs period the 25 V gap reduces to 37% of 25 V $=9{\cdot}25$ V. Thus $V_c=109$ V and $V_r=(V_b-V_c)=(100-109)$ V $=-9$ V.




(a)




(b)

Fig. 11.9(a) Showing the waveform produced by a 40 μs field pulse partial-differentiator circuit on 405-line even and odd fields. The waveforms differ only slightly at the beginning of the field pulse train, but much more so at the end.

Fig. 11.9(b) The waveform shown in Fig. 11.9(a) appears like this if viewed on an oscilloscope with the X-timebase running at a slow speed (e.g. 10 ms/cm). When the trace is expanded by turning up the X-gain control the waveform shown in Fig. 11.9(a) resolves itself.

At $t_{12}$ — $V_b$ rises to 180 V. At this instant $V_r=(V_b-V_c)=(180-109)$ V$=71$ V. Thus the leading edge of the fourth half-line pulse rides up to 71 V positive to chassis.

Between $t_{12}$ and $t_{13}$ — $C$ charges from 109 V towards 180 V. In the 10 μs period the 71 gap narrows by $0{\cdot}2\times 71$ V$=14$ V. Thus $V_c=(109+14)$ V$=123$ V and $V_r=(V_b-V_c)=(180-123)$ V $=57$ V.

At $t_{13}$ — $V_b$ drops to 100 V. At this instant $V_r=(V_b-V_c)=(100-123)$ V$=-23$ V.

Between $t_{13}$ and $t_{14}$ — $C$ discharges from 123 V down towards 100 V. In the 40 μs period the 23 V gap closes to 37% of 23 V$=8{\cdot}5$ V.

At $t_{14}$ — $V_b$ rises to 180 V. At this instant $V_r=(V_b-V_c)=(180-108{\cdot}5)$ V$=71{\cdot}5$ V.

The fifth half-line pulse hence rises to $=+71{\cdot}5$ V and differs by only 0·5 V from the fourth half-line pulse which reaches +71 V. The waveform has therefore reached a state of equilibrium, with equal areas either side of the zero voltage line. The remaining half-line pulses level off at 71·5 V.

At $t_{22}$ the signal changes its pattern and $V_b$ rises to 180 V at a moment when $V_r$ is at $-18$ V; i.e. $V_c$ is at 118 V. Thus $V_r=(V_b-V_c)=(180-118)$ V$=+62$ V. In the 90 μs period between $t_{22}$ and $t_{25}$, $C$ charges to 174 V and $V_r=(180-174)$ V$=+6$ V.

When a similar procedure is adopted to trace the sequence of events on the pulse train which follows odd fields, the differentiated waveform is found to differ from that of the even field in the following respects:

(1) The first half-line pulse rides up some 4 V higher than the corresponding even field pulse. This difference in amplitudes of the half-line pulses on alternate fields quickly diminishes so that the waveforms become coincident at about the third half-line pulse.

(2) A substantial difference occurs at the trailing edge. The reason for this is self-evident from a study of the applied signals in this region.

With this circuit it is the half-line pulses (which separate the true field sync. pulses), which are utilised for synchronisation of the field oscillator. Usually the timebase is made to trigger on the first of these pulses. The complete waveform is clipped above the level reached by the line pulses; e.g. along the line XY shown in Fig. 11.8(d) and Fig. 11.9(a). The way in which this is achieved is described in Chapter 13.

It should be noted that the leading edges of the half-line pulses are coincident on odd and even fields. They are also vertical so that the interval between the start of the first radiated field pulse and the leading edge of the derived field sync. pulse is the same on both fields. This appears to give the field pulse differentiator a distinct advantage over the integrator—a point which will be more clearly understood after the following chapter on Interlacing has been studied. However, the circuit has two less obvious disadvantages as a result of which it has found less favour than the integrator. These are:

(a) The effect on interlacing of the substantial waveform differences on the trailing edge of the field pulse sequence. This matter is taken up in the next chapter.

(b) The circuit is fast acting and produces vertical output pulses for every vertical voltage change seen by it. Thus it responds very readily to interference pulses whereas the slow-acting integrator absorbs and rides out all but the worst of such impulsive interference.

*The partially differentiated 625-line field pulse train*

Fig. 11.10 illustrates the effect of a field pulse partial differentiator on the 625-line waveform. As with the integrator diagrams, the value of the equalising pulses is clearly brought out.

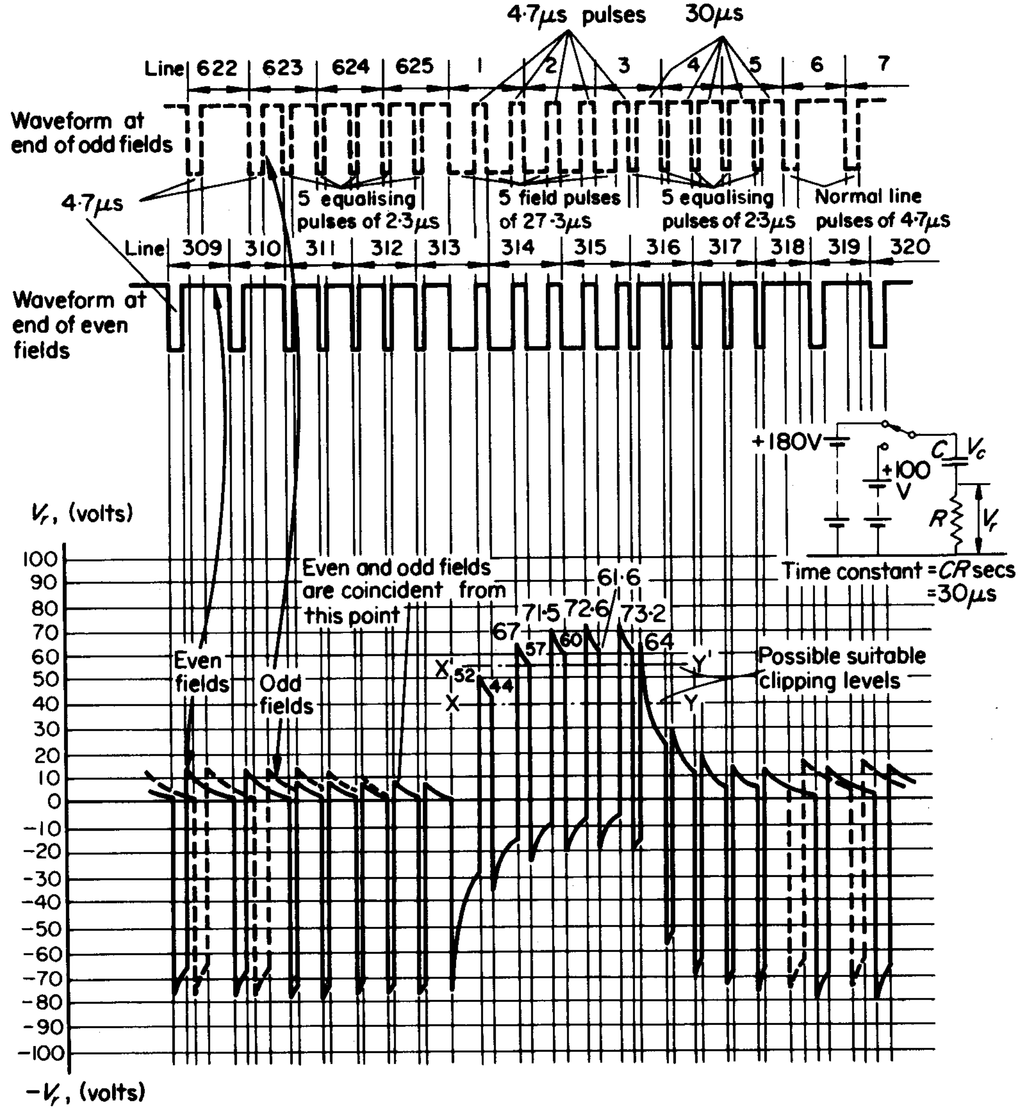



Fig. 11.10 Showing the action of a 30 μs field pulse partial-differentiator circuit on a British 625-line waveform. The presence of equalising pulses before and after the field sync. pulses results in identical waveforms being produced across $R$ during the actual field picture sequence.

Their presence is seen to remove the discrepancies between odd and even fields. The waveform is clipped above the voltage level reached during line sync. pulses (e.g. along the line XY), so that the output waveforms reaching the field oscillator are identical on both fields.

If the 625-line signal used positive modulation this would now make the field pulse differentiator a most attractive proposition. However, its susceptibility to interference still remains, and with negative modulation it is likely to be exposed to heavier interference, although this is not a serious problem on u.h.f. transmissions.

# 12

# Interlacing

There are certain precise requirements which, if satisfied, *must* give rise to perfect interlace, in all field timebase circuits. With some timebase oscillator circuits it is possible to achieve good interlace (usually by luck rather than design) without meeting these requirements. However, it is as well to start by looking at the ideal situation, and then to see how far it is possible to depart from this in certain specific practical circumstances.

**Requirements for perfect interlacing with all types of oscillators**

(1) The field pulses fed to the vertical oscillator must be identical in every respect on odd and even fields.

(2) The time interval between the start of the radiated field pulse chain and the leading edge of the derived field sync. pulse in the receiver must be exactly the same after odd and even fields.

Since the radiated field pulse chain is inserted regularly twice per picture, and the picture consists of an odd number of lines, it is obvious that if these conditions are met the resulting raster must be interlaced.

At this stage it is helpful to summarise some basic facts about a television raster. To begin with, the following points are worth remembering:

(a) Every part of the sawtooth oscillatory cycle of the field oscillator represents a particular vertical level on the screen.

(b) If it were possible to halt the field oscillator at a particular point in its cycle, the line oscillator would cause the spot to draw a horizontal line at that appropriate level.

(c) During any normal picture line the field oscillator is slowly lowering the spot throughout the line. Thus the line slopes gently downwards from left to right.

(d) During the line flyback period, the downward movement is much less because the line flyback time is much shorter than the line scanning time. Hence the flyback line is almost horizontal, sloping only very gradually downwards from right to left.

(e) During the field flyback the spot is being lifted upwards whilst the line timebase sweeps it from side to side. The flyback lines slope upwards from left to right. The slope is steeper than the downward slope of the picture lines because the field flyback stroke is of course much faster than the field scanning stroke.

(f) There is a particular horizontal level at the top of the screen which corresponds to the level reached by the spot at the end of the field sawtooth flyback stroke. This level must represent the starting level for the scanning stroke of the spot, on both odd and even fields.

(g) Similarly there is another level at the bottom of the screen which is reached at the end of the scanning stroke. This level is the level from which the flyback movements start on both odd and even fields.

(h) The flyback time; i.e. the time it takes the spot to move from the lower level (g) to the upper level (f) is precisely the same on odd and even fields. (This is always true if the field pulses are identical as in requirement 1 above. Under certain circumstances it can also be true even if this condition is not met.)

It follows from the last three points that, if the starting points of the flyback journeys are half a line apart on odd and even fields, then the finishing points (hence the starting points of the next succeeding fields) are also separated by a distance of half a line.

Suppose that the starting point of the downward journey of the spot at the start of one field is at the extreme left of the screen as shown in Fig. 12.1(a). This point is shown to lie on the horizontal line AB which represents the level referred to under (f) above. At the end of the first line the spot has moved a certain distance downwards; the slope of the line being exaggerated in the diagram to illustrate the point clearly.

The starting point of the next field is at point C, in the middle of the screen. Point C again lies on the horizontal line AB. The vertical distance the spot has dropped by the end of the half a line which represents the start of the new field is clearly only one-half of the distance dropped by the full-length first line of the previous field.

On return to the left of the screen the spot must arrive at a level midway between the first two lines of the previous field. Thereafter all successive lines of this field interlace between those of the previous field.

In a first approach to the study of interlacing, an idealised raster is considered, as was done in Chapter 1 (see Fig. 1.2(b)).

Such a raster involves two impossible conditions.

(A) That the field timebase is triggered precisely at the centre of the last line of the first field and at the end of the last line of the second field.

(B) That flyback is instantaneous so that none of the total lines per picture is lost during flyback.

The next step, which is still hypothetical, but which moves nearer to being realistic, acknowledges the impossibility of the condition (B) but retains condition (A). Thus we may move further towards an understanding of an interlaced raster by making the following propositions:

(i) That the field timebase is triggered by the leading edge of the first field sync. pulse after each field. This now establishes the starting point of the field flyback at the precise centre of the last line of one field and at the extreme end of the last line of the alternate field. This establishes condition (A) above.

(ii) That the actual flyback lasts from this moment of triggering to the end of the last suppressed line of the field suppression period.

If these conditions are fulfilled then it is obvious that the first *active* line of one field starts at the top left-hand edge of the raster and the first active half a line of the other field starts at the top of the raster midway across the screen (Fig. 12.1(a)).

Proposition (i) cannot be true in practice, however, since if the field timebase were able to respond to the *leading* edge of the *first* field pulse, then it could equally well respond to the leading edge of line pulses. Near the end of its scanning stroke a timebase is in a critical condition and inevitably the field oscillator would be in danger of being triggered by the last line pulses occurring just before the first field pulse.

Proposition (ii) could conceivably be true, but there is no vital necessity why it should be.

This is so because once the field timebase has embarked upon its flyback excursion, a certain precise time will elapse before it completes it. This time is entirely controlled by the receiver circuit constants. There is nothing in the radiated waveform which dictates to the timebase that flyback is over and that the new scanning stroke should commence. The flyback time

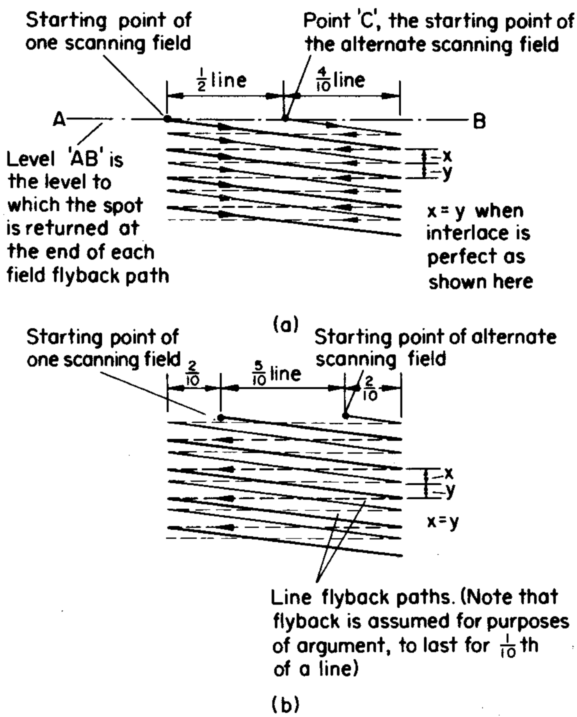



Fig. 12.1 Showing examples of perfect interlace

Points to note:

(i) In each case the starting points of the two scanning fields are separated by a time equal to half a line; e.g. by 50 μs with the 405-line signal, or 32 μs on 625-lines.

(ii) If the field flyback times are equal on both fields, the starting points of the two field flyback paths are also separated by exactly half a line. (The argument in the text should be followed with care.)

(iii) The dimensions shown (e.g. 1/2 line, 2/10 line, etc.) are *time* dimensions in terms of the time duration of one complete video signal line (i.e. fractions of 100 μs for the 405-line signal, or of 64 μs for 625-lines).

(iv) In the examples shown in (a) and (b), line flyback is assumed for purposes of argument, to occupy exactly 1/10 of a line; i.e. 10 μs on 405 or 6·4 μs on 625. The dimensions shown along AB, which refer to the length of the scanning raster lines, hence add up to 9/10 in each case.

(v) Since line flyback time is a constant, and is not an issue which affects interlace, in the interlace error diagrams of Fig. 12.2 line flyback is for the sake of simplicity taken as being instantaneous, so that on those diagrams the dimensions along AB add up to unity.

can therefore occupy a time which is less than, equal to, or greater than, the part of the field suppression period which still remains after the moment of triggering.

In the first case, some of the post field pulse suppressed lines are scanned by the line timebase while the spot is on its way up the screen and the remainder form the first few lines of the new field. This is the normal condition found in an efficient receiver. In the second case the end of

flyback exactly coincides with the end of the last suppressed line and all field lines subsequently scanned are active lines. This corresponds to proposition (ii) above. In the last case, flyback has not ended when the first active line begins. This implies that 'fold-over' must occur because the first lines of the next field are traced by the spot as it completes its movement to the top of the screen. This last case is clearly undesirable and indicates a fault condition.

It is important to note that if the first proposition were true, correct interlace would still result in all three of the cases mentioned. This follows because the durations of the flyback movements on odd and even fields are assumed the same. Thus whatever the time taken for the spot to reach the top of the screen, since the flybacks originate from points half a line apart, they must terminate at points which are also half a line apart.

It remains therefore to examine the effect of modifying the first proposition to establish what actually happens in practice.

Let us now assume that the *original* two requirements (1) and (2) for perfect interlace are safely met. The variable quantities from one receiver to another are now:

(a) The time interval between the start of the radiated field pulse chain and the moment when the timebase triggers. In any particular receiver we have stipulated that this interval is to be equal on odd and even fields, but the *actual* interval involved will vary from receiver to receiver.

(b) The time duration of the flyback excursion of the field timebase. This is dictated by the receiver circuit constants.

It is clear that endless slightly different permutations of these variables are possible. In all cases, however, interlace will be theoretically perfect. The question of length of flyback time has been looked at and its effects noted. If a specific arbitrary length for this is stipulated the effect of variations in the first variable quantity, i.e. the time interval mentioned under (a) above, may be examined.

Thus let the flyback occupy a time equivalent to 12 complete lines. Suppose, for the purpose of argument, that the critical field oscillator triggering voltage level is reached in a time interval equivalent to three-tenths of the time duration of one line after the leading edge of the first field pulse (i.e. 30 μs after the leading edge of the first field pulse with the British 405-line waveform and 19·2 μs with the corresponding 625-line waveform).

It must at this point be remembered that after one of the two fields for each picture it is the leading edge of the first field pulse of the chain which triggers the line timebase to start the first line of the next field (see the first field pulse following the ends of lines 625 and 405 of the respective signals shown at the top of Figs 11.10 and 11.9).

Thus, at the end of one complete picture, the spot is at the bottom right-hand corner of the screen when the first field pulse arrives. In the idealised raster it would now move instantaneously to the top left-hand corner of the screen. In practice, for the case considered, the spot would fly over to the bottom left-hand corner and start moving back across the screen. The question arises: Where exactly will the spot be situated when the field flyback excursion starts?

Suppose a time equal to one-tenth of a line is allowed for line-flyback. In the suggested interval of three-tenths of a line between the leading edge of the first field pulse and the moment the field timebase triggers, the spot will have moved to the extreme left of the screen and will be a distance equivalent in time duration to two-tenths of a line (i.e. 20 μs for the 405-line signal or 12·8 μs for the 625-line signal) along a new line (i.e. along line 1) when the field timebase triggers. For convenience of argument, the field flyback has been said to be a whole number of lines (i.e. 12) in duration. This implies that the spot reaches the top of the screen on line 13,

two-tenths of a line from the left. The particular line concerned, in this specific case, is one of the suppressed lines (see line 13 of the odd field of the 405-line waveform and line 13 of the even field of the 625-line waveform shown in Figs 3.4 and 3.5). The first part (two-tenths) of this line is scanned by the spot as it completes the very last part of its upward movement, and the downward scanning stroke for the new field hence starts by the completion of the remaining portion of this 13th line. The situation is illustrated in Fig. 12.1(b).

At the end of this field, the first field pulse occurs half-way through a line, i.e. half-way between two successive leading edges of line sync. pulses, the leading edge of a line pulse serving as the usual reference time for measurement purposes. A time of one-tenth of a line has been allowed in this argument for line flyback. Thus, in stating that the field pulse arrives half-way through a line, it is implied that it starts five-tenths of a line after the last line sync. pulse. But the first one-tenth of a line corresponds to line flyback, so that the leading edge of the first field pulse is now seen to occur at a moment when the spot is a distance, equal in time, to four-tenths of a line from the left-hand edge of the screen.

Allowing the same time interval as before, of three-tenths of a line between the leading edge of the first field pulse and the firing of the field timebase, this now fixes the position of the start of the field flyback at a point $\frac{4}{10}+\frac{3}{10}=\frac{7}{10}$ of a line from the left-hand edge of the screen. The flyback will terminate 12 lines later at the top of the screen at a point seven-tenths of a line from the left, and the new scanning excursion starts from here. Since on the previous field the starting point was two-tenths of a line from the left of the screen, the two fields start at points exactly half a line apart and hence interlace is again assured.

It is interesting to note that the requirement for perfect interlace is that the starting points of successive fields must be half a line apart. There is no reason why these two points should be the extreme left and top dead centre of the screen. Notwithstanding this, for convenience and simplicity, in the cases of errors of interlace now to be dealt with, it is assumed that the starting point of one field is always at the extreme top left-hand edge of the screen. This allows for easy representation of interlace errors by the simple expedient of moving the other field's starting point about, on either side of its then correct position, one half-line to the right.

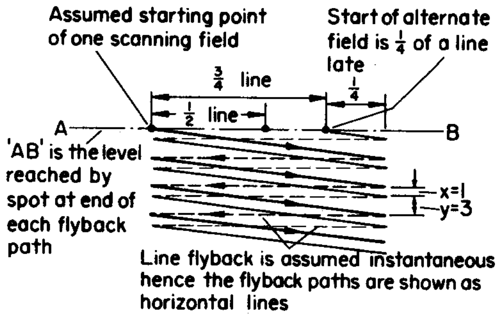
It will also be noticed that the dimensions along the line AB of Fig. 12.1 add up to nine-tenths of a line. The other one-tenth of a line, as explained, has been allowed to take account of line flyback. This tends to complicate the reasoning and in the diagrams which follow, showing examples of interlace errors, i.e. Fig. 12.2(a), (b), (c), (d) and (e), line flyback has been assumed to be instantaneous. Since the line flyback time is constant, this is not an issue which affects interlace. The proportions along the line AB in these sketches would need to be modified slightly to take account of line flyback, but the principle remains the same.

The diagrams are self-explanatory and each should be examined very carefully. The points to be borne in mind when studying them are summarised below.

(i) For simplicity, line flyback is here assumed instantaneous.

(ii) For convenience one scanning field is assumed to start from the extreme left-hand edge of the raster. This allows interlace errors to be illustrated by moving the starting point of the alternate field by various amounts either side of its then correct central position.

(iii) If the field flyback times are equal, the errors may be related to the starting points of the field flyback traces. This in turn allows the cause of the interlace error to be identified in terms of differences in the shape and timing of the derived composite field sync. pulses on alternate fields. As examples of this, 'Case A' of Fig. 12.3 and Fig. 12.4 should be correlated with 12.2(d) and 'Case B' with 12.2(b), when the appropriate text has been studied.




(a)

(a) 50% interlace error

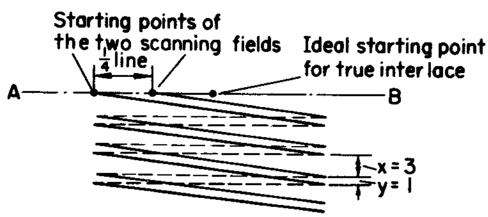
The alternate scanning field is $\frac{1}{4}$ line late in starting. This causes very poor interlace, showing up as line pairing. The inter-line spaces are in the ratio 1:3, giving 25:75 interlace instead of 50:50.

*On 405* Alternate scanning field starts 25 μs late, i.e. starting points are separated by 75 μs instead of 50 μs.

$$\% \text{ interlace error} = \frac{75-50}{50} \times 100 = 50\%$$

*On 625* Alternate field starts 16 μs late. Starting points are 48 μs apart instead of 32 μs.

$$\% \text{ interlace error} = \frac{48-32}{32} \times 100 = 50\%$$




(b)

(b) 50% interlace error

Alternate field is $\frac{1}{4}$ line early in starting. This causes the same degree of pairing as (a), i.e. the ratio of inter-line spaces is 3:1, giving 75:25 interlace.

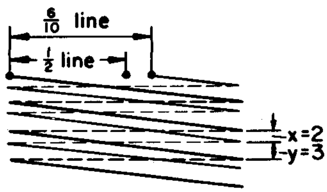
*On 405* Alternate field starts 25 μs early, i.e. starting points are separated by 25 μs instead of 50 μs.

$$\% \text{ interlace error} = \frac{50-25}{50} \times 100 = 50\%$$

*On 625* Alternate field starts 16 μs early. Starting points are 16 μs apart instead of 32 μs.

$$\% \text{ interlace error} = \frac{32-16}{32} \times 100 = 50\%$$

Fig. 12.2 Showing examples of interlace errors




(c)

(c) 20% interlace error

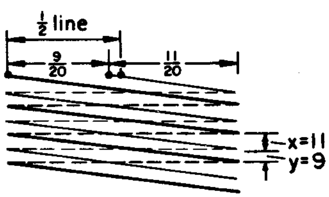
Alternate field is $\frac{1}{10}$ line late starting. Inter-line spaces are in the ratio 2:3, giving 40:60 interlace.

*On 405* Alternate field starts 10 μs late. Starting points are separated by 60 μs instead of 50 μs.

$$\% \text{ interlace error} = \frac{60-50}{50} \times 100 = 20\%$$

*On 625* Alternate field starts 6·4 μs late. Starting points are 38·4 μs apart instead of 32 μs.

$$\% \text{ interlace error} = \frac{38{\cdot}4-32}{32} \times 100 = 20\%$$




(d)

(d) 10% interlace error

Alternate field is $\frac{1}{20}$ line early starting. Inter-line spaces are in the ratio 11:9, giving 55:45 interlace. This ratio is sometimes quoted as being a reasonable design tolerance.

*On 405* Alternate field starts 5 μs early. Starting points are separated by 45 μs instead of 50 μs.

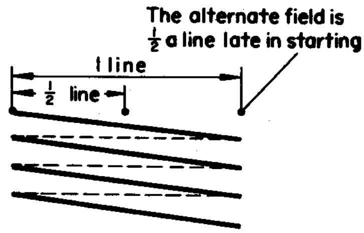
$$\% \text{ interlace error} = \frac{50-45}{50} \times 100 = 10\%$$

*On 625* Alternate field starts 3·2 μs early. Starting points are separated by 28·8 μs instead of 32 μs.

$$\% \text{ interlace error} = \frac{32-28{\cdot}8}{32} \times 100 = 10\%$$

Fig. 12.2 Showing examples of interlace error

Fig. 12.2 (cont.)




(e)

(e) 100% interlace error

Alternate field is $\frac{1}{2}$ line late starting. This results in the scanning lines of the two fields being superimposed one upon another, giving the appearance of a rigid heavy line structure, i.e. there is a complete absence of interlace.

*On 405* Alternate field starts 50 μs late. Starting points are separated by a complete line of 100 μs.

$$\% \text{ interlace error} = \frac{100-50}{50} \times 100 = 100\%$$

*On 625* Alternate field starts 32 μs late. Starting points are separated by 64 μs instead of 32 μs.

$$\% \text{ interlace error} = \frac{64-32}{32} \times 100 = 100\%$$

*Interlace errors due to pulse differences*

A certain critical pulse amplitude is needed to trigger a timebase oscillator. The value of this amplitude varies from circuit to circuit, whilst in a particular circuit it varies with the setting of the frequency (hold) control.

Fig. 12.3 illustrates the point. A blocking oscillator valve is cut off during the scanning period and highly conductive during flyback. The diagrams show the grid voltage waveform. The oscillator is assumed to be unsynchronised and working at its free-running frequency. For synchronisation to be possible, the free-running frequency has to be lower than the required field frequency. The diagrams allow us to see what sync. pulse amplitude is needed to trigger the oscillator. The field frequency is 50 c/s, so that one complete field oscillator waveform lasts for 20 ms. Flyback is initiated at the moment when the decaying voltage across the grid auto-bias capacitor crosses the grid cut-off level. The separation between the leading edges of successive flyback strokes must therefore be 20 ms in the locked condition.

The diagram of Fig. 12.3(a) shows the pulse amplitude needed in one arbitrary case if the oscillator is to trigger 20 ms after the leading edge of the last flyback stroke. In Fig. 12.3(b) the free-running frequency is assumed to be lower still which is seen to necessitate an increase in the required sync. pulse amplitude.

An important point must be made here. The sync. pulse amplitude available is made large enough to trigger the oscillator over quite a range of adjustment of the hold control, but adjustment of the hold control when the picture is locked *must* still change the actual amplitude of the sync. pulse *needed* to lock the oscillator. This can have repercussions on interlace, and cause the degree of interlace to change with adjustment of the field hold control, sometimes from almost perfect interlace to severe line pairing. This is obviously undesirable since the user will normally merely twiddle the hold control until the picture stays still, and the degree of interlace achieved is then a mere matter of chance.

This raises again the question of disparity between field sync. pulse leading edges on odd and even fields. The integrator has been shown to produce a pulse having a sloping leading edge; i.e. the sync. pulse amplitude increases with time. It follows that the bigger the sync. pulse needed by an oscillator the longer is the interval between the leading edge of the first radiated

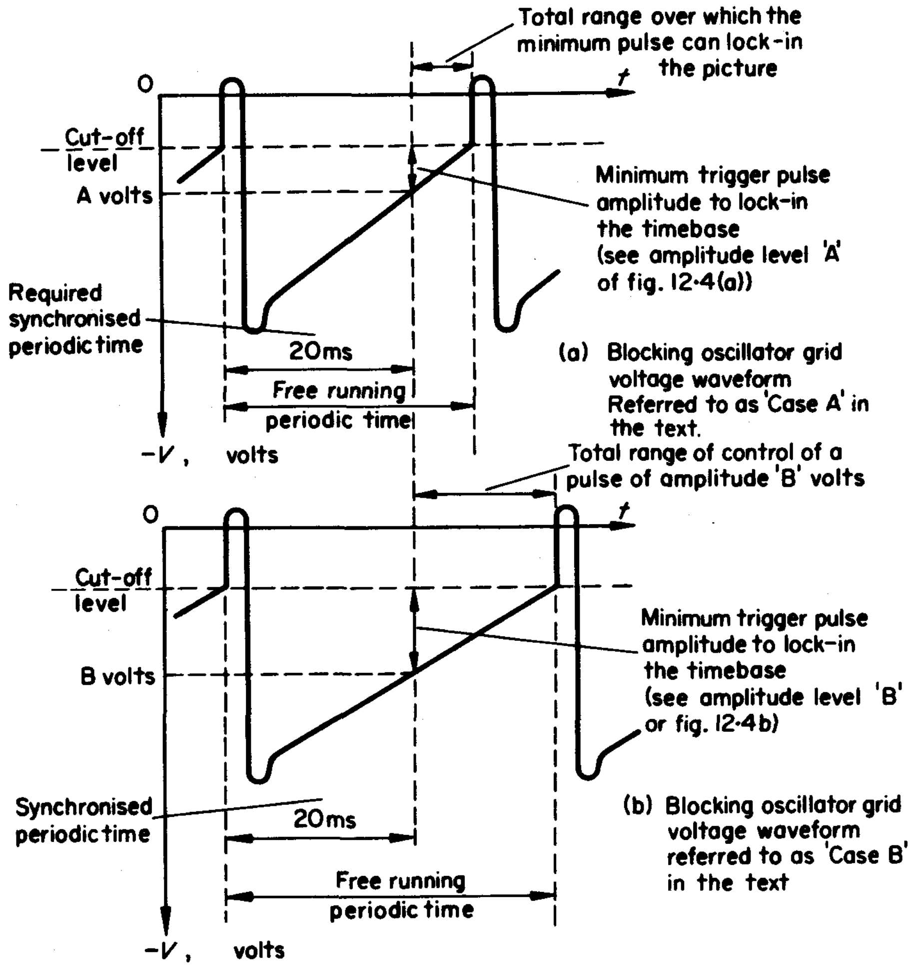



Fig. 12.3 Showing that, when the free-running speed of a timebase oscillator is changed, the necessary minimum trigger pulse amplitude changes

Reference to Fig. 12.4 shows how it is that the degree of interlace can change when the field hold control is adjusted; e.g. a change of free-running speed may increase the necessary minimum pulse amplitude from A to B, which, in turn, may affect the quality of interlace.

field pulse and the moment of triggering. This in itself is of no consequence if odd and even field pulses are identical. When an integrator is used on television waveforms which do not include equalisation pulses, however, the edges are *not* coincident, as was clearly shown in Fig. 11.6. The effect of this is examined in Fig. 12.4 which refers to the 405-line signal.

These diagrams are related both to the oscillator waveforms of Fig. 12.3 and the interlace error diagrams of Fig. 12.2.

Suppose the hold control is adjusted so that the free-running frequency would be as shown in 'Case A' of Fig. 12.3(a). The necessary sync. pulse amplitude is shown as 'A' volts. The oscillator will trigger when the integrated waveform reaches this amplitude. But Fig. 12.4(a) shows that the amplitude level of 'A' volts is reached in a shorter time after odd fields than it is after even fields. If the derived field pulse following the even field is chosen as the reference,

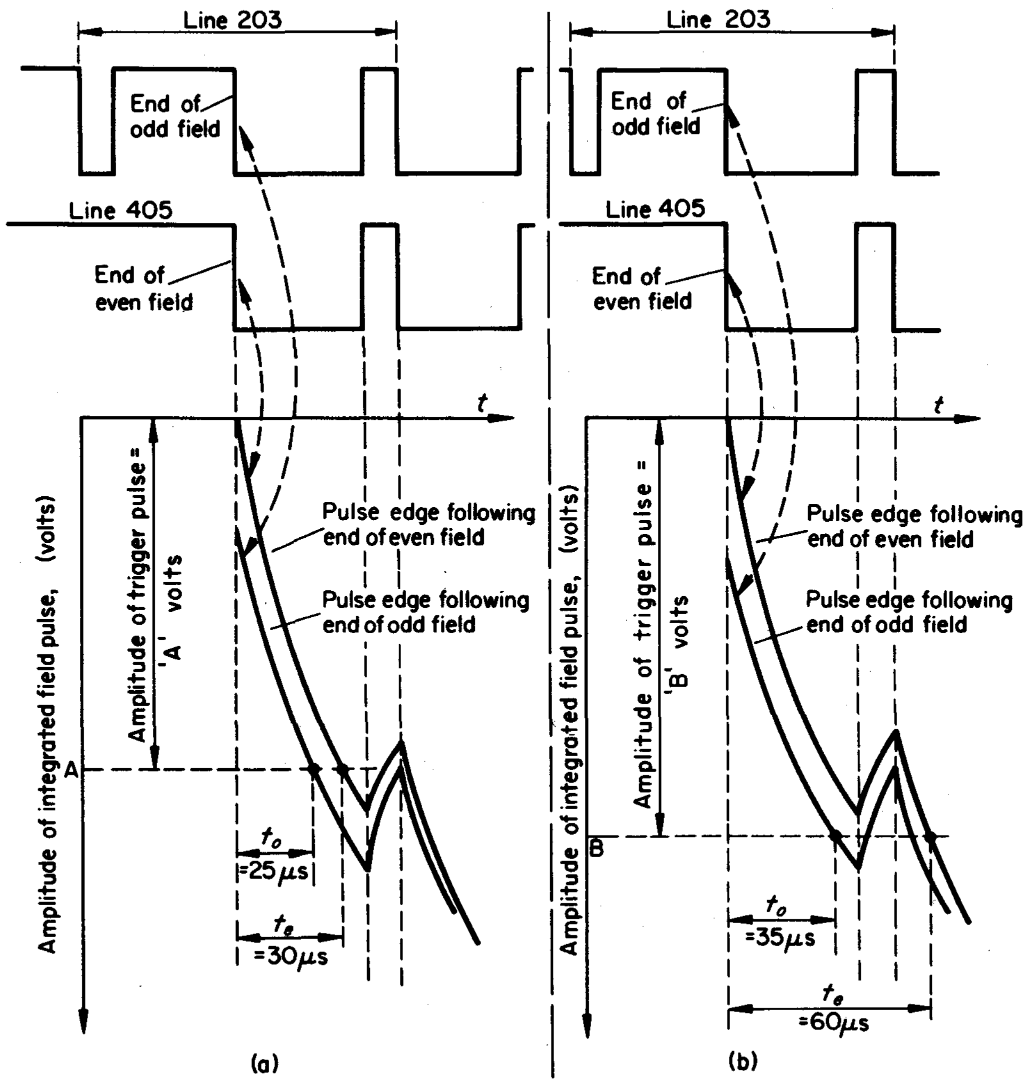



Fig. 12.4 Showing how a disparity on the leading edge of the integrated field pulse impairs the quality of interlace

(a) Following even fields the field timebase triggers $t_e = 30$ μs after the start of the pulse train. Following odd fields the interval is $t_o = 25$ μs; i.e. the timebase triggers $(t_e - t_o) = 5$ μs early. This represents an interlace error of $5/50 \times 100 = 10\%$. (See Fig. 12.2(d).)

(b) With this triggering level the timebase fires on the first field pulse following odd fields but on the second following even fields. The disparity is $(t_e - t_o) = (60 - 35) = 25$ μs. This represents an interlace error of $(25/50) \times 100 = 50\%$. (See Fig. 12.2(b).)

it may be said that after odd fields the oscillator triggers early; in this particular case 5 μs early. This represents an interlace error of 10%; a figure sometimes quoted as a reasonable design tolerance. This degree of interlace corresponds to Fig. 12.2(d).

Now suppose the hold control is adjusted so that the free-running frequency would be that shown in 'Case B' of Fig. 12.3(b). Reference to Fig. 12.4(b) shows that an entirely different situation now exists. After even fields, pulse amplitude 'B' is not reached until 10 μs of the *second* field pulse has elapsed, but after odd fields the same amplitude is reached just before the end of the *first* pulse. In this instance, therefore, after odd fields the oscillator triggers 25 μs early. This causes an interlace error of 50% with severe line pairing, as is illustrated in Fig. 12.2(b).

Trailing edge differences in the pulse shapes on odd and even fields can also cause interlace troubles. Even if the leading edges are coincident, if the trailing edges are not, interlace errors can still appear. In this particular circumstance, which, as shown in the previous chapter, is characteristic of the pulse waveform produced by field pulse partial differentiators, examination of the field flyback lines on a receiver shows perfect interlace of the half-line pulses. This indicates that the field oscillator is being triggered correctly and one might expect to find interlace on scan. Often, however, this is not so. The difference in the pulse shapes at the trailing edges causes a difference in the total flyback time of the two fields and hence the starting times of the scanning strokes are incorrectly spaced.

The energy content of the sync. pulse fed to the field oscillator is proportional to the area of the pulse. A change of shape represents a change of energy content, and this affects the time-constant networks which control the frequency of the oscillator. The sync. pulse does not therefore cease to be of interest after the moment on its leading edge when flyback is initiated. It is combinations of capacitors and resistors which determine the frequency of timebase sawtooth generators. The capacitors charge and discharge through resistors and the performance of the circuit must be affected by all the electrical influences impinging upon it. Just how much effect the trailing edge discrepancy has varies from circuit to circuit and is difficult to predict.

It is instructive to examine the discussed field by field discrepancies in the timing and/or width of the receiver's derived field sync. pulses, against the background of the 20 ms (i.e. 20,000 μs) periodic time of the field timebase itself.

As demonstrated in Fig. 12.2(e) an error of half a line in the start of a scanning stroke gives rise to complete line pairing, i.e. to 100% interlace failure. This represents errors of 34 μs and 50 μs in the 625-line and 405-line signals respectively. An error of one quarter of a line (i.e. of 16 μs on 625-lines or 25 μs on 405-lines) gives a 50% interlace error, as is shown in (a) and (b) of Fig. 12.2. Similarly an error of one tenth of a line (i.e. 6·4 μs on 625-lines and 10 μs on 405-lines) gives rise to a 20% interlace error.

To speak of errors of 'half a line' and a 'quarter of a line', with a design target of an error not exceeding one twentieth of a line, does not in itself put the matter truly into perspective. When these errors are expressed as ratios against the periodic time of the field oscillator, however, it becomes easier to see why some care is needed in the production at the receiver of the derived field sync. pulse. Table 12.1 shows examples of interlace errors with the corresponding timing errors shown as percentages of the periodic time.

As pointed out, the errors in the starting times of the scanning strokes on alternate fields may be due to the flyback strokes starting at unequal intervals, or to the flyback times being of slightly different duration, or to a combination of both of these factors.

The presence of pre-field-sync.-pulse equalising pulses in a television signal protects the receiver against the possibility of out-of-step timing of the start of flyback on alternate fields,

whilst the post-field-sync.-pulse equalising pulses protect it against the possibility that small—but highly significant—differences will exist between the actual duration of the flyback strokes on alternate fields.

**Table 12.1**

*Interlace errors and the corresponding error-time/periodic-time ratios*

| Interlace error | 625-line signal | | 405-line signal | |
|---|---|---|---|---|
| | Alternate field timing error | Ratio of timing error to periodic time | Alternate field timing error | Ratio of timing error to periodic time |
| 10% | 3·2 μs | $\frac{3{\cdot}2}{20{,}000} = 0{\cdot}016\%$ | 5 μs | $\frac{5}{20{,}000} = 0{\cdot}025\%$ |
| 20% | 6·4 μs | $\frac{6{\cdot}4}{20{,}000} = 0{\cdot}032\%$ | 10 μs | $\frac{10}{20{,}000} = 0{\cdot}05\%$ |
| 50% | 16 μs | $\frac{16}{20{,}000} = 0{\cdot}08\%$ | 25 μs | $\frac{25}{20{,}000} = 0{\cdot}125\%$ |
| 100% | 32 μs | $\frac{32}{20{,}000} = 0{\cdot}16\%$ | 50 μs | $\frac{50}{20{,}000} = 0{\cdot}25\%$ |

As inferred earlier, receivers working on television signals which do not include equalising pulses in the field synchronising waveform, may still be made to give very good interlace but require more care in the design of the field oscillator synchronising arrangements and are often (but not always) very prone to changes in interlace quality with changes in the setting of the field-hold control over its lock-in range.

Finally it must be stressed that a receiver field oscillator is a simple circuit which certainly does not call for a high degree of engineering accuracy. It is simply a slave of the incoming signal and the transmitter sets the ultimate standards of timing and accuracy. Given regularly spaced identical *derived field pulses*, the field oscillator is *compelled* to produce good interlace over the whole of its lock-in range. Without them it can hardly be expected to do so when it is realised what wide differences of interlace quality can result from pulse discrepancies which are so small compared with the oscillator's periodic time.

It has been seen that it is possible to have perfect interlace on flyback but poor interlace on scan. Also possible, though less common, is good scanning interlace accompanied by poor interlace on flyback. This can arise when both leading and trailing edge discrepancies exist because in some circumstances these can be partly compensatory. The leading edge discrepancy causes alternate fields to trigger at points not quite separated by half a line so that the half-line pulses seen on the flyback pattern show evidence of pairing. The trailing-edge discrepancy may now partly correct this error by causing the total flyback time to vary slightly from one field to the next.

Of course, failure to interlace on flyback is not in itself of any consequence. The matter is

discussed to draw attention to the fact that a study of field flyback lines is certainly no guide to the state of scanning interlace. However, it is instructive to examine the pattern of flyback lines which appear when the brightness control is turned up to make black-level visible, and the field flyback suppression circuit is disconnected from the tube. It allows us to picture the path of the spot from bottom to top of the screen and to identify the video-signal lines of the field blanking period. The pattern is best seen when a dim picture is being radiated or during the brief pause between one programme and another or one scene and another. The half-line pulses between the field pulses, and the suppressed lines which follow the field pulse sequence, show up brighter than the rest of the raster because these areas of the screen phosphor are scanned and illuminated twice; once during the scanning stroke and again during flyback. The actual field pulses are at minimum video signal voltage level and so of course they do not show up. The half-line pulses and the suppressed lines carry the video voltage level up to 30% of maximum on the 405-line and 23% on the 625-line signals, so that these are seen.

The pattern for the 405-line signal is shown in Fig. 12.5 and the lines are related by letters to the parts of the signal causing them.

The diagram shown is reproduced from a tracing taken from the screen of a receiver in which the field pulses are separated by a field pulse partial differentiating circuit. The field timebase triggers on the leading edge of the first half-line pulse of each field; i.e. on the trailing edge of the first broad field pulse.

At the end of line 405 the line timebase triggers on the leading edge of the first field pulse and the spot sweeps to the left of the screen and begins a new line. By the time the first field pulse ends, the spot has progressed some distance back across the screen, tracing out part of what is actually line 1 of the next odd field. The field timebase triggers on the trailing edge of this pulse and the spot begins to lift. The 10 μs top of the half-line pulse which follows this first field pulse is seen on the screen as a short bright line. This pulse is marked 'A' on the diagram. Similarly the last 10 μs of the line is brightened up because the video signal returns to suppression level for this period (see B). On the leading edge of the third field pulse the line-timebase triggers again and the spot sweeps back to the left of the screen. Meanwhile the flyback stroke of the field timebase is lifting the spot back towards the top of the screen and its progress upwards may be traced by studying the lines one by one.

It is interesting to note that the trailing edge of the first field pulse at the end of odd fields occurs just 10 μs before the end of line 203. Hence, since as was stated above, the field timebase is triggered by the trailing edge of the first field pulse, the spot does not begin to lift until just before the end of line 203 and the flat top of the pulse 'P' is not seen since it is usually beneath the tube mask. In the idealised raster the spot is assumed to start its journey back to the top in the middle of line 203 after odd fields, and at the end of line 405 after even fields. In the practical case described above, triggering takes place near the *end* of line 203 and towards the *middle* of line 1. This reinforces the observation made earlier that the points of departure of the spot must be half a line apart for interlace to be perfect, but the precise positions along a line of these points is unimportant. On a receiver having a faster flyback than the one shown, the columns of short lines will extend further up the screen and the post-field pulse suppressed lines will be more closely packed at the top of the screen.

A matter which must be mentioned in passing is that one of the suppressed lines near the end of the field blanking period is sometimes used to carry special test signals (such as the pulse and bar signal referred to in Chapter 6, or perhaps a 'staircase' signal), which are used for testing the performance of various sections of the television transmission system. Since these signals carry the modulation level up above black level, often to peak white, it follows that

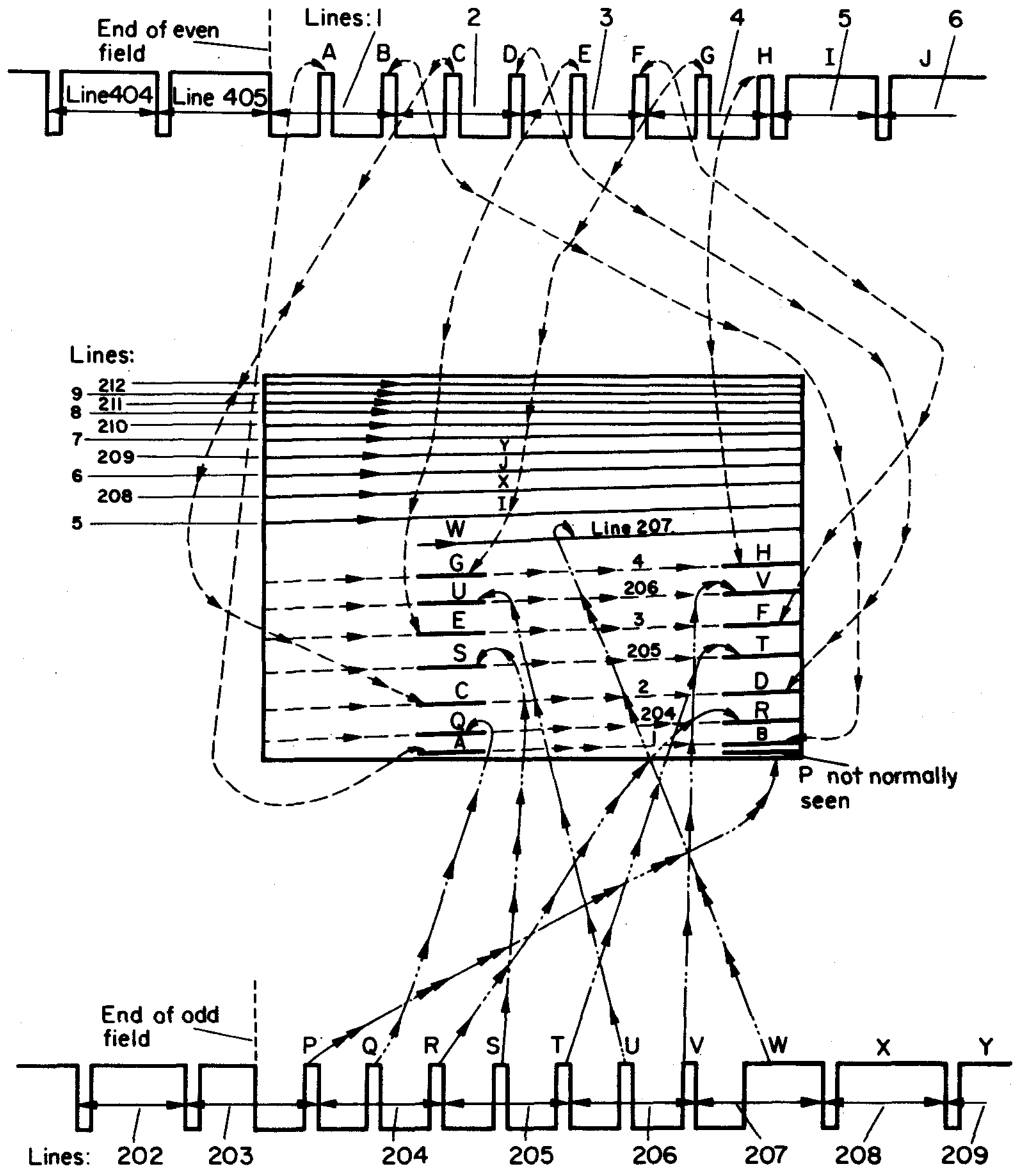



Fig. 12.5 Pattern of flyback lines seen on a 405-line raster with the brightness control turned up to make black-level visible, and the field flyback suppression circuit disconnected

they will be visible on the screen of a receiver in which the field flyback has not finished when they occur. On receivers with a fast field flyback these lines produce no detrimental effect on the picture, but in some instances, where flyback is slower, the information is seen near the top of the screen as bright sections of lines sloping upwards from left to right. The signals are present on both fields and therefore appear twice, close together; again showing evidence of interlace on flyback. Where these signals prove a nuisance it is usually a simple matter to

modify the field oscillator to speed up flyback and so shift the offending lines up under the top edge of the screen mask.

Judging the quality of interlace from the screen is often not easy and requires experience. It must be remembered that although there are 50 fields per second, each individual line only reappears at the rate of 25 times per second. There is therefore line flicker on a bright picture. If complete pairing is taking place then a particular line on the screen is being drawn 50 times per second and no flicker appears. The raster takes on a rigid steady appearance. An interlaced raster creates the impression of continuous line movement. This makes it difficult to discern how good the interlace is by gazing at the screen.

A useful method of examining quality of interlace is to view a small section of the raster through a narrow horizontal slot (e.g. 2 in $\times \frac{1}{4}$ in) cut in a postcard. The card is held to the screen face and the line structure in the small window is then clearly visible. On some receivers it is interesting to adjust the field hold control over its lock-in range whilst studying the interlace. For the reasons described, wide differences in interlace quality at various settings will sometimes be noticed.

A final point must be made clear. With most receivers, field flyback finishes before the end of the suppressed lines, so that the next field starts with the remaining suppressed lines. These lines are not seen since, if the receiver is set up correctly, the spot is not visible when the modulation rests at blanking level. As soon as picture modulation starts, however, the position of the scanning spot may be identified. Since modulation starts at the beginning of one line on one field and at the centre of a line on the alternate field, the *visible* raster starts at these points. It should be noted that these points mark the start of picture detail and *not* the start of the scanning fields. The actual starting points of the two fields are a little higher up and usually lie under the tube mask (as indeed do the first lines of the active picture itself in most receivers).

However, notwithstanding this difficulty in identifying the actual starting points of the two fields, if the first visible indication of the raster shows one line starting about half-way across the screen, positioned clearly just above the next line down on the raster, then the indication is that some degree of interlace *is* being achieved. The actual degree of interlace can only be judged as indicated in Figs 12.1 and 12.2, by examining the ratio of adjacent inter-line spaces.

# 13

# Field Pulse Processing Circuitry

The action of line-pulse differentiators and field-pulse integrators has been discussed. The circuit of Fig. 13.1 shows an example of the appearance of these basic elements in a commercial receiver.

The resistor $R_2$ forms part of the anode load from which the line pulse differentiator is fed, and is also the resistive element of the field pulse integrator. It must be remembered that it is

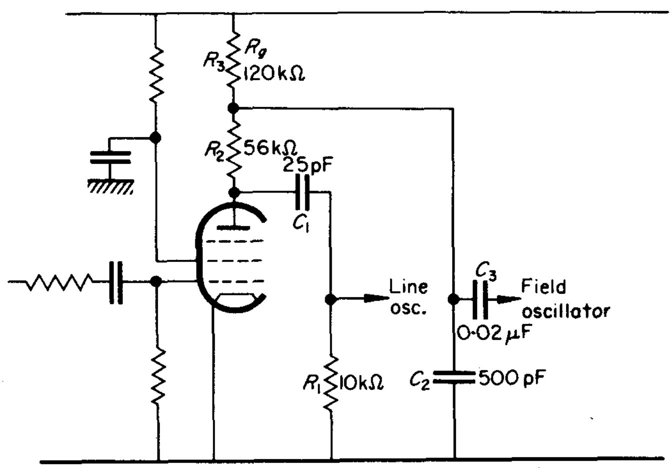



Fig. 13.1 Example of use of simple differentiator and integrator in a sync. separator circuit. The associated time-constants are as follows:

(i) Time-constant of differentiator $= C_1R_1$ seconds

$$= 25 \times 10^{-12} \times 10 \times 10^3 \times 10^6 \text{ μs}$$
$$= 0{\cdot}25 \text{ μs}$$

(ii) Time-constant of integrator when $C_2$ is discharging (i.e. during sync. pulses) $= C_2R_2$ seconds

$$= 500 \times 10^{-12} \times 56 \times 10^3 \times 10^6 \text{ μs}$$
$$= 5 \times 5{\cdot}6 \text{ μs}$$
$$= 28{\cdot}0 \text{ μs}$$

(iii) Time-constant of integrator when $C_2$ is charging (i.e. between sync. pulses) is

$$C_2R_3 \text{ seconds} = 500 \times 10^{-12} \times 120 \times 10^3 \times 10^6 \text{ μs}$$
$$= 5 \times 12 \text{ μs}$$
$$= 60 \text{ μs}$$

the negative-going movement of potential across the integrator capacitor $C_2$ which forms the derived composite field sync. pulse. Between the incoming sync. pulses, when the sync. separator valve is cut-off, $C_2$ charges towards the H.T. potential via $R_3$. During a sync. pulse, when the valve is highly conductive, the lower end of $R_2$ is effectively connected to chassis through the low d.c. resistance of the valve. $C_2$ then discharges towards chassis potential through $R_2$ and the conductive valve. It follows that the charge and discharge time constants are different. However, the important time-constant is the one governing the rate of discharge of $C_2$ and this, as is shown on the diagram, is 28 µs. The re-charging time constant is 60 µs.

In a practical circuit the re-charging time-constant is always longer than the charging time-constant. The reason for this may be revised at this point by reference back to the conventional circuit shown in Fig. 11.1. As discussed in Chapter 11, the anode load resistor forms part of the total series resistance through which the capacitor charges when the valve is off, but conversely it does not affect the discharge time when the valve is on. Obviously, the same is true of both the differentiator and integrator in Fig. 11.1.

It has been seen to be desirable to clip off the integrated field pulse beyond the level reached by the line pulses, i.e. along the line XY of Fig. 11.7. Circuits employed to clip the sync. pulse waveform in this way appear under a variety of names, e.g. clippers, limiters, interlace diodes, etc.

## D.C. coupled single-diode clippers

The principle of a simple field pulse clipper is shown in Fig. 13.2. By means of the potential divider $R_4$, $R_5$, the diode anode is held at a positive potential which is less than that of the

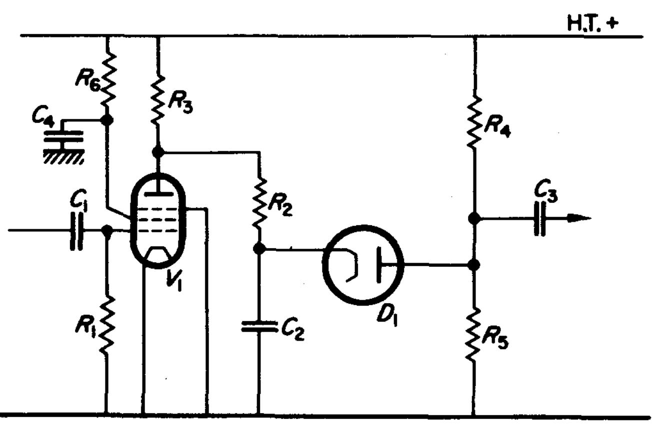



Fig. 13.2 Circuit showing the basic principle of a diode frame pulse 'clipper' or 'limiter'

cathode at all times except during field pulses. The diode is thus normally cut off and open-circuits the path from the sync. separator to the field oscillator.

$R_2$ and $C_2$ form an integrator. The waveform across $C_2$ is of the form shown in Figs 11.6 and 11.7. The negative-going movement of the voltage across $C_2$ during line pulses is insufficient to drop the diode's cathode potential below its anode potential. Such pulses are therefore ignored by the diode. During the field pulses, however, the voltage across $C_2$ sweeps downwards on the leading edge of the integrated waveform and the diode's cathode is driven negative to its anode. The diode conducts and a negative-going pulse appears across $R_4$. This is passed

on to the field oscillator via the d.c. blocking capacitor $C_3$. The diode stops conducting when the trailing edge voltage level is such that its cathode is once again positive to the anode. The integrated waveform is therefore sliced off. The slicing level may be altered by adjustment of the potential divider $R_4$, $R_5$.

Fig. 13.3 shows a practical example of this form of clipper.

The positive voltage for the anode of limiter diode $D_1$ is obtained by connecting the anode directly to the field oscillator anode. This is fed through a 330 kΩ resistor so that during normal

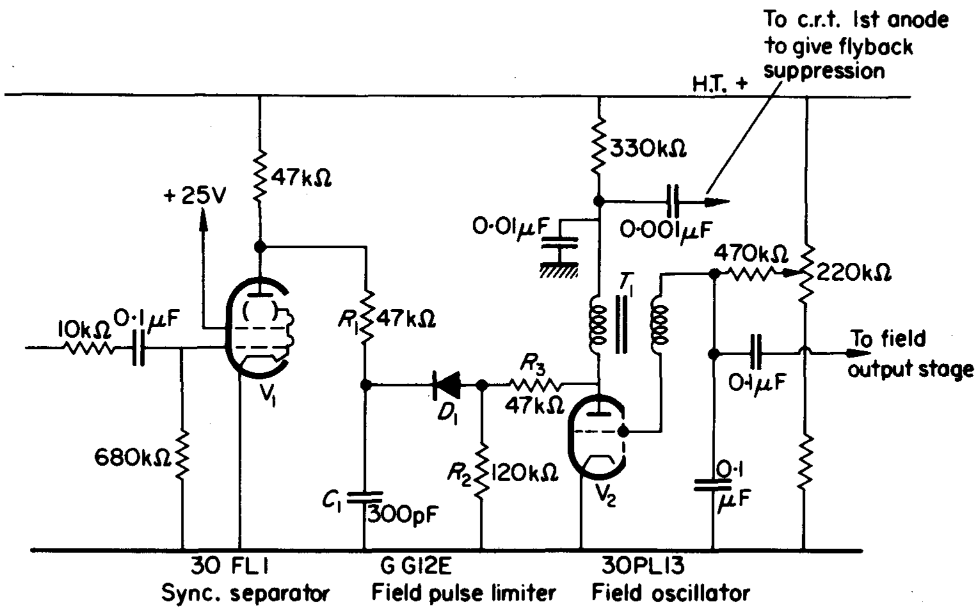



Fig. 13.3 Example of a circuit using the principle of Fig. 13.2

picture lines the diode anode is considerably lower in potential than the cathode, which returns to the sync. separator anode. $R_1$ and $C_1$ form the integrator, and the operation of the circuit is as described for Fig. 13.2.

A point of interest arises in the manner of application of the sync. pulse to the field oscillator. The latter is a blocking oscillator which requires a positive-going sync. pulse on its grid to trigger it.

The pulse output from $D_1$ is, however, negative-going. This negative pulse is applied to the oscillator anode and coupled by the blocking oscillator transformer to the grid circuit, where it appears as a positive-going pulse. Obviously, since the circuit is an oscillator, the transformer is connected to produce 180 degree-phase shift in the anode to grid feedback path, i.e. to give positive feedback. The transformer hence works as a sync. pulse invertor; a device which is often used.

### A.C. coupled diode clippers

The circuit of Fig. 13.4 works in a rather different way. Both the anode and cathode of the diode are returned to chassis.

The field pulses from the sync. separator are taken from the pentode screen grid instead of anode. This gives the advantage of separating the field and line timebase circuits which normally find a meeting point at the sync. separator anode, with the resulting possibility of interaction. When the sync. separator is driven into conduction during sync. pulses, the screen current through $R_4$ produces a negative-going voltage change at the screen, in the same way as the anode current through $R_1$ causes a negative-going movement of the anode voltage. $R_2$ and $C_2$, however, in the lower arm of the screen potential divider, form a long time-constant circuit

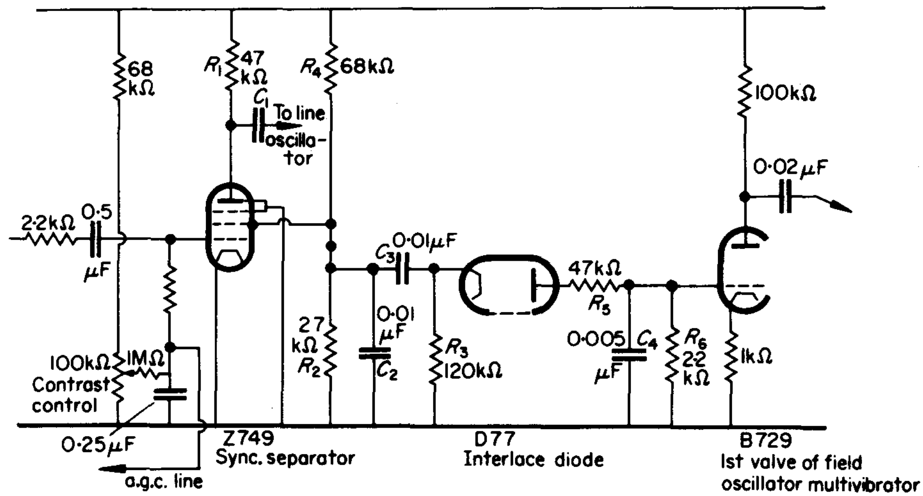



Fig. 13.4 Interlace diode circuit (a.c. coupled)

(270 μs). This causes the negative-going voltage movement at the screen to be very small during the brief line pulses. $C_3$ charges through $R_3$ to the screen potential, but the time-constant of $C_3R_3$ is very long (1200 μs), and the small voltage changes across $C_2$ due to line pulses cause no significant current flow through $R_3$. The diode cathode is thus at chassis potential, as is the anode, and the diode is non-conductive.

During the field pulse chain, the screen is conductive for successive field pulses and only cut off for the brief periods of the half-line pulses. The voltage across $R_2C_2$ now has time to change and moves less positive to chassis. This causes $C_3$ to begin to discharge towards the new potential facing it. Electrons leave the 'right-hand' plate of $C_3$ and travel down through $R_3$ to chassis. The top of $R_3$ is driven negative to chassis so that the diode's cathode becomes negative to its anode. The diode conducts causing a negative-going sync. pulse to appear at the oscillator grid across $C_4R_6$.

At the end of the field pulse sequence, the screen potential rises again, and $C_3$ begins to charge to this higher voltage. Electron current up through $R_3$ causes the diode cathode to be driven sharply positive so that the diode cuts off.

A further example of this a.c. coupled type of 'interlace diode' is shown in Fig. 13.5. The sync. pulse waveform from the sync. separator anode is directly coupled to the grid of a valve whose chief purpose is to produce anti-phase line sync. pulses for a flywheel synchronisation circuit. An integrator $R_1C_1$ is connected between the anode of this triode and earth. The time-constant of this circuit is $(1200 \times 10^{-12} \times 33 \times 10^3 \times 10^6)$ μs $= 39{\cdot}6$ μs. The integrated waveform

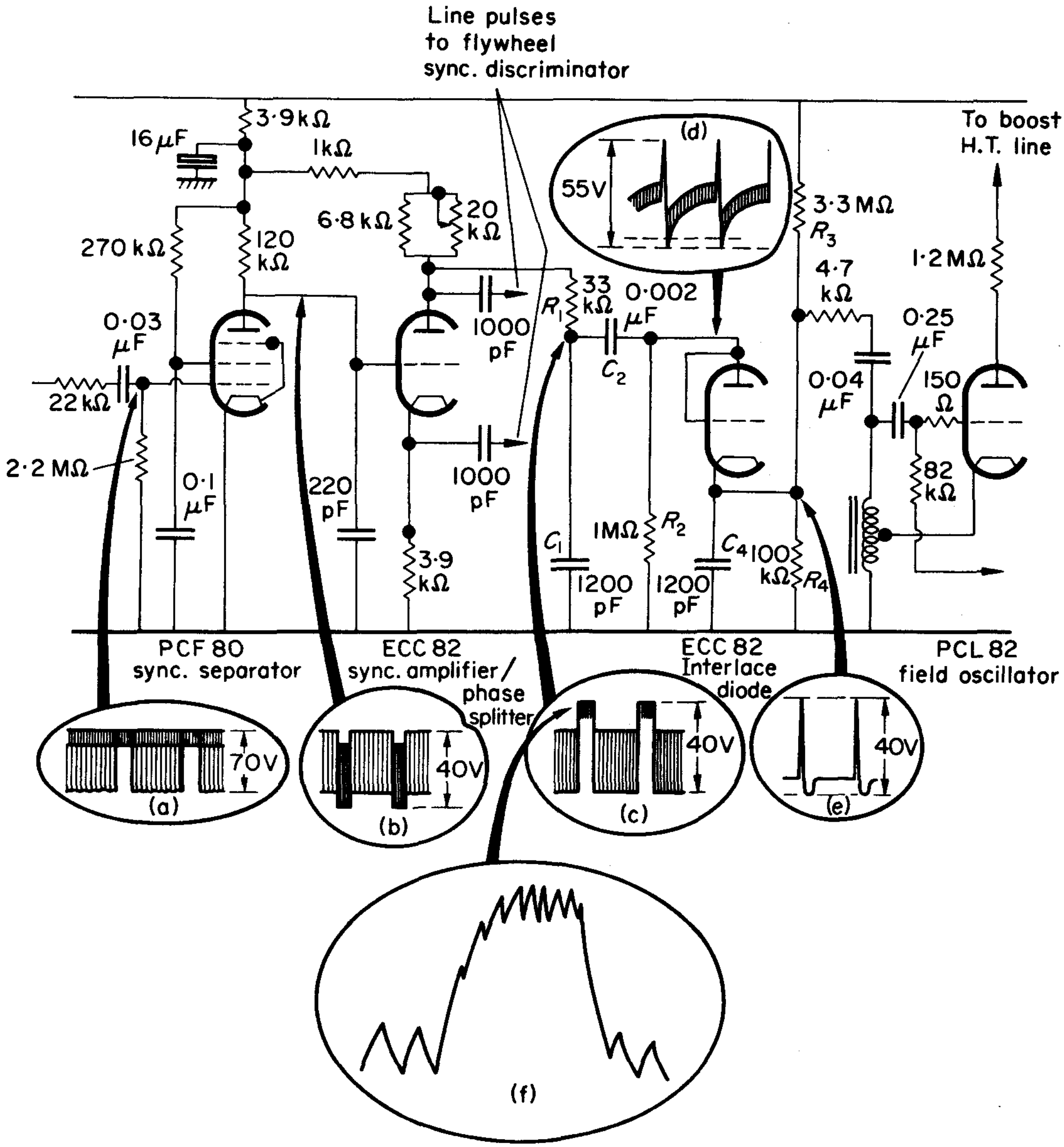



Fig. 13.5 Interlace diode following a sync. amplifier/'phase splitter' (405-line receiver)

*Note.* The waveforms are those which would be seen with the oscilloscope X-timebase running at approximately picture speed. It is obvious that no precise detail can be observed, but the position of the two fields and the field pulses is evident.

produced across $C_1$ is passed to the interlace diode anode via a differentiator $C_2R_2$ which has a longer time constant of $(0{\cdot}002\times10^{-6}\times1\times10^{6}\times10^{6})$ μs = 2000 μs. The diode cathode is held positive to chassis by the potential divider $R_3R_4$. It must be noted that the integrated waveform at the triode anode is positive-going; i.e. the inverse of the waveform at the sync. separator anode.

At the triode anode the voltage level is therefore such that it rises during sync. pulses and falls during the interim periods. The rise of voltage across $C_1$ during line pulses is only small. $C_2$ charges through $R_2$ to the voltage across $C_1$, but again the long time-constant prevents the current flow through $R_2$ from being very significant during line pulses. During the successive field pulses there is a much larger positive-going movement across $C_1$, and this is differentiated by $C_2R_2$ to give a positive pulse of some 20 V amplitude at the diode anode. This exceeds the cathode voltage so that the diode conducts. The positive-going pulse at the cathode is passed on via a stand-off resistor and a d.c. blocking capacitor to the field blocking oscillator grid.

The diagram shows sketches of the waveforms at various points as seen on an oscilloscope running at a speed comparable to that of the picture (e.g. 10 ms per cm). Such waveforms give a misleading impression of the sharpness of the voltage changes during field pulses. To put the matter in perspective, diagram (f) shows the integrated waveform across $C_1$ as it would be seen if the trace were expanded out.

It may be wondered why the waveform (d) is able to fall rapidly from its positive peak yet rises very slowly following the negative peak. This is due to the influence of the diode. During the positive peak the diode is conductive so that $C_2$ now charges via this low resistance path instead of through $R_2$. The cathode capacitor $C_4$ provides the necessary reservoir of current to give a sharp pulse. When the trailing edge of the integrated waveform across $C_1$ begins, the diode is still conductive until the voltage falls below its clipping level. The diode then cuts off and the discharge of $C_2$ is now governed by the 2000 μs time constant of $C_2$ and $R_2$.

The final active field pulse produced by this circuit, shown at (e), is free from line pulses and is of some 40 V amplitude.

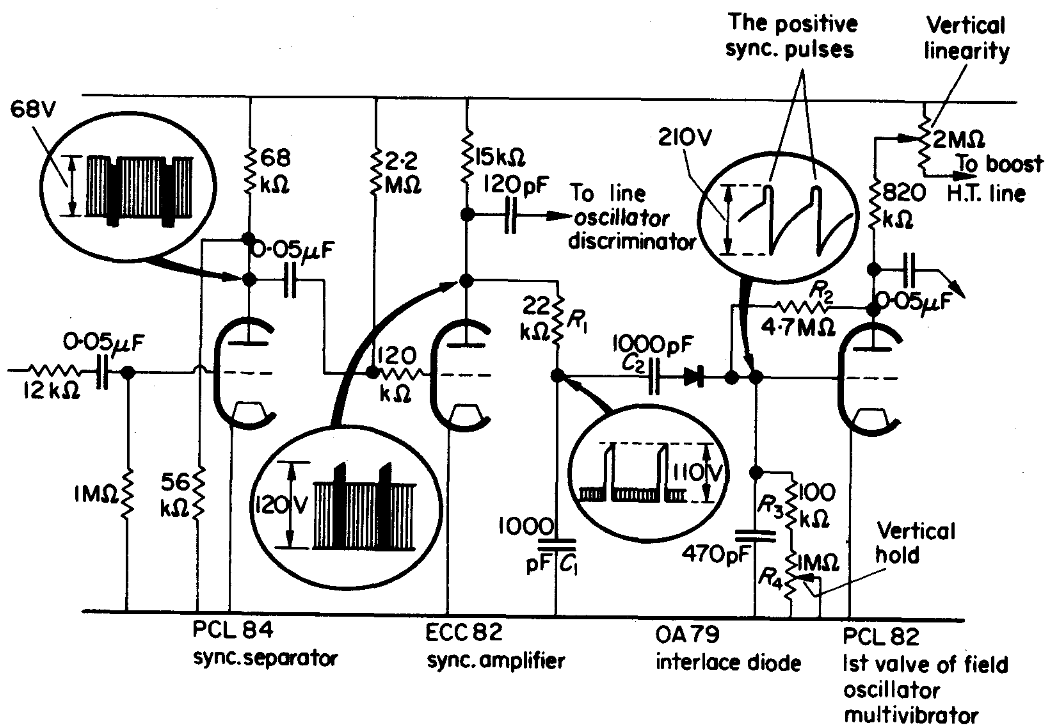



Fig. 13.6 Similar circuit to Fig. 13.5, as used in a 625-line receiver

Fig. 13.6 is a similar circuit used in a 625-line receiver. The triode is, in this case, a.c. coupled to the sync. separator and functions as a straightforward sync. pulse amplifier. A small positive potential is maintained on the control grid which ensures that between sync. pulses the anode current is saturated. The valve is then only able to respond to negative-going changes at the grid and produces clean positive-going pulses at its anode. $R_1$ and $C_1$ form an integrator having a time constant of $(1000 \times 10^{-12} \times 22 \times 10^3 \times 10^6) = 22$ µs. As in the previous circuit, a positive-going integrated field pulse waveform appears across $C_1$.

This waveform is passed via $C_2$ to the anode of the interlace diode. The cathode of the diode is held positive to chassis by a potential divider $R_2$, $R_3$, $R_4$ connected between the field oscillator anode and chassis. When the circuit is first switched on, the diode conducts and $C_2$ charges through $R_1$, and the 15 kΩ anode load resistor to the same potential as that reached by $C_1$. Once $C_1$ has charged the diode cuts off. On the arrival of the field pulse signal, the positive-going voltage across $C_1$ carries the diode anode positive to its cathode, so that it conducts to allow $C_2$ to follow the voltage change across $C_1$. The electron current up through $R_4$ and $R_3$ drives the grid of the triode section of the field multivibrator positive, so triggering the valve into conduction and initiating the flyback stroke.

If there were a resistor between the diode side of $C_2$ and chassis, it would be easier to see how it is that the diode anode is driven positive. Thus as $C_2$ charged through this resistor to follow the increasing potential across $C_1$, the electron flow would drive the top of the resistor positive with respect to chassis; i.e. drive the diode anode resistor positive. However, the semiconductor diode is not an open circuit even when reversed biased, so that there is in fact a resistive path between $C_2$ and chassis. The forward bias across the diode is therefore produced as before. At the end of the field pulse, when the diode cuts off, $C_2$ gradually discharges once again, through the reverse resistance of the diode.

**The triode clipper**

Yet another version of a clipper or limiter circuit is shown in Fig. 13.7. As indicated earlier there is nothing highly critical about component values and time constants in these circuits. This particular one serves both the 405- and 625-line signals with equal efficiency. A triode clipper (here called an 'interlace filter'), is used. The grid is held at a positive potential by returning the grid resistor $R_2$ to the H.T.+ line, so that the valve passes continuous grid current. A high-value anode load resistor is employed, and between field sync. pulses the anode voltage is very low.

$R_1$ and $C_1$ form an integrator of time-constant $(82 \times 10^{-12} \times 330 \times 10^3 \times 10^6) = 27$ µs. The integrator is fed from the anode of the sync. separator and therefore develops a negative-going waveform. $C_2$ must follow the voltage changes on $C_1$ so that during sync. pulses, when the voltage across $C_1$ falls the net charge on $C_2$ must rise. Electrons flow from $C_2$, up through $R_2$, producing a negative-going voltage change across this resistor. The drop in voltage across $C_1$ during line pulses is small, and the resulting negative-going change of potential across $R_2$ is insufficient to overcome the effect of the positive voltage on the grid of the triode. The triode anode current remains saturated and no voltage change takes place at the anode.

During the integrated field pulse waveform, however, the much larger drop in voltage across $C_1$ results in a greater charging current to $C_2$. The electron current up through $R_2$ now drives the triode grid negative causing a sharp drop in anode current, with a corresponding positive-going change at the anode. This positive pulse is fed via the d.c. isolating capacitor $C_3$ to the blocking oscillator grid.

The waveform (c) is a combination of the sync. pulse and the oscillator grid voltage wave-

form. The waveforms (a) show that the peak-to-peak voltage changes across $C_1$ are almost the same on the 405- and 625-line signals, but the line pulses are larger in the former case because their longer duration results in a larger voltage change from the 27 μs integrator.

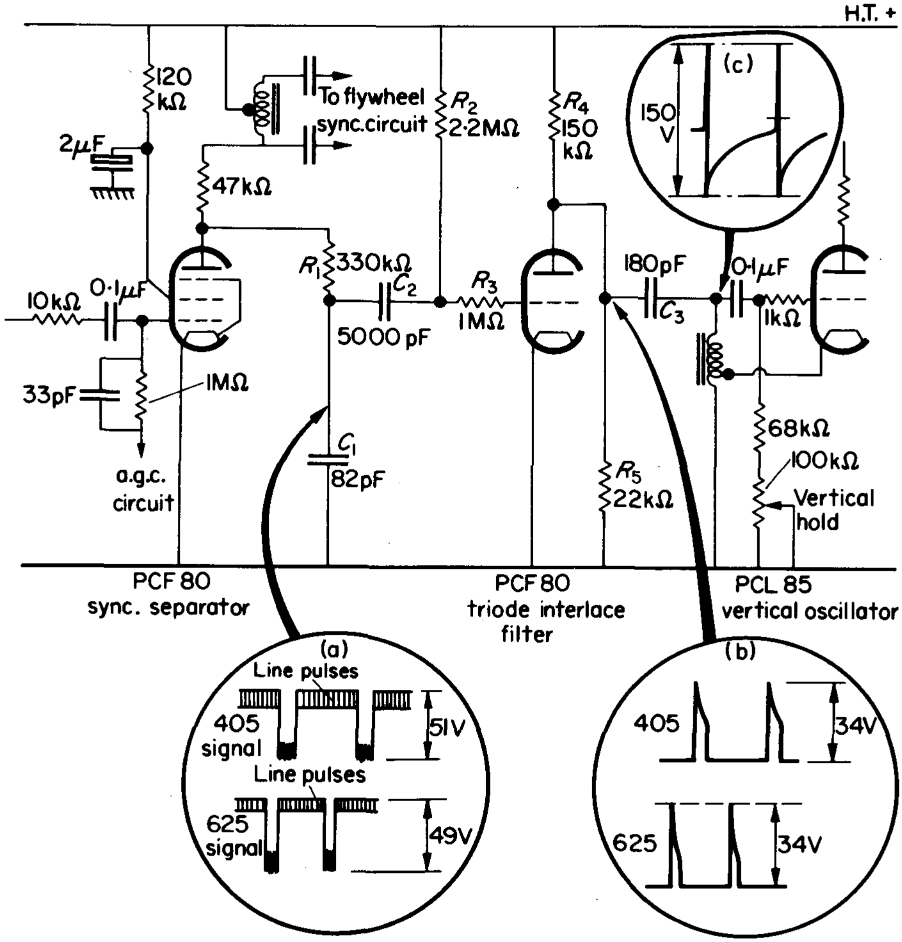



Fig. 13.7 Showing a triode interlace filter in a dual-standards 405/625 receiver

### Double-diode interlace filter

The double-diode interlace filter of Fig. 13.8 employs a different principle. Between sync. pulses the sync. separator is cut off and its anode rests at the full H.T. potential. $D_1$ is conductive and the cathode capacitor $C_2$ charges to a potential which is approximately that which would be present across $R_2$, if $R_1$ and $R_2$ were simply a potential divider connected between the H.T. line and chassis. In the instance shown, this places the voltage across $C_2$ at roughly 180 V. $D_1$ passes a small current all the time and the cathode voltage remains constant.

The cathode of $D_2$ is strapped to that of $D_1$. This results in $D_2$ being cut off since *its* anode potential, which is fixed by the potential divider $R_3R_4$, is arranged to be below that of the cathode. If the H.T. voltage is 200 V the anode of $D_2$ is placed at approximately 120 V positive to chassis; i.e. at a level which is some 60 V negative to the cathode.

The arrival of a line sync. pulse causes the sync. separator anode voltage to drop sharply. This carries $D_1$ anode voltage below that of the cathode and $D_1$ is cut off for the duration of the pulse. $C_2$ now discharges through $R_2$ and the cathode voltage falls. The network $R_2C_2$ has a long time-constant, however, so that the fall of voltage during the line pulse is limited. The circuit is arranged so that this drop in voltage is insufficient to carry the voltage at $D_2$ cathode

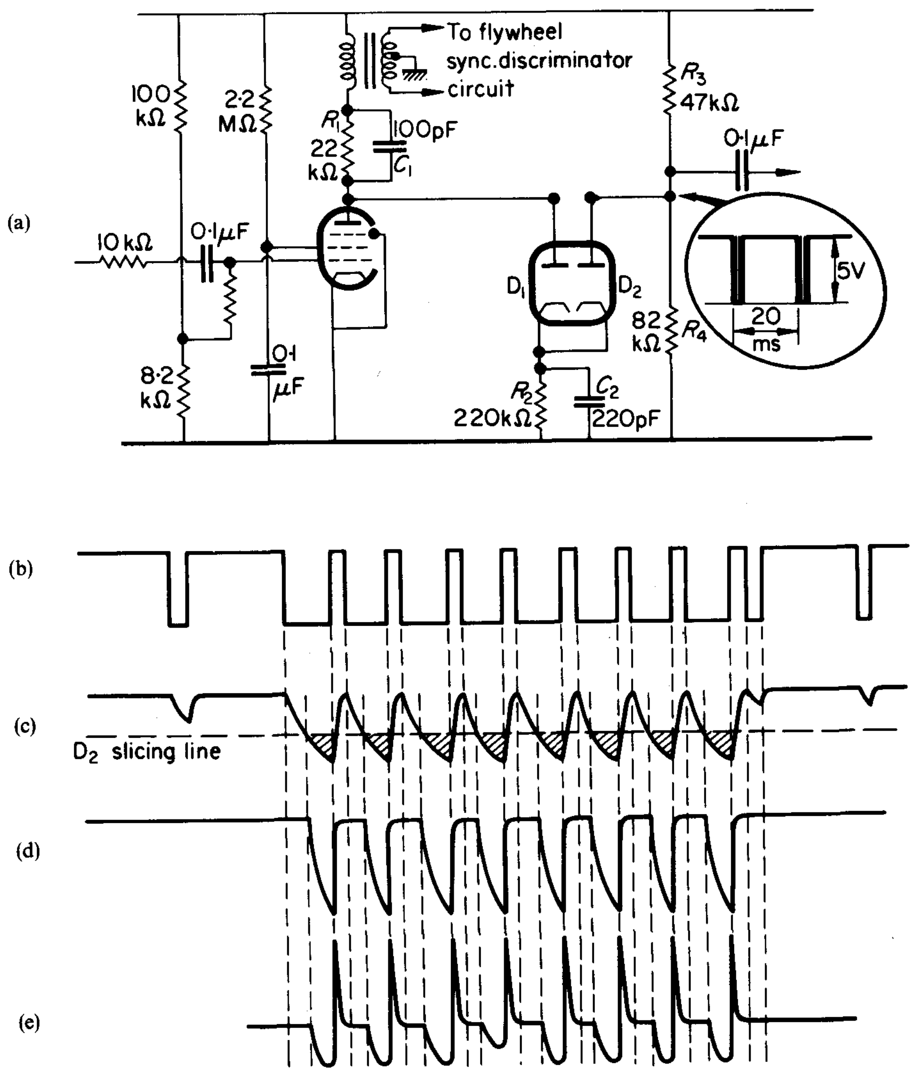



Fig. 13.8(a) Double-diode interlace filter circuit

Fig. 13.8(b) Waveform at anode of $D_1$

Fig. 13.8(c) Waveform at $D_1D_2$ cathode

Fig. 13.8(d) Waveform at anode of $D_2$

Fig. 13.8(e) The effect of differentiating the anode waveform of $D_2$

below that of the anode. A time-constant of the same order as the time duration of one field pulse is a useful figure for such circuits. The circuit shown was developed for 405-line working. The cathode time constant is 48 μs.

During a field pulse $C_2$ is able to discharge much further, and before the end of such a pulse the cathode voltage falls below that at $D_2$ anode. $D_2$ therefore begins to conduct and a negative going output pulse appears at $D_2$ anode. When the field pulse ends, the sync. separator anode potential rises sharply for the duration of the half-line pulse. This immediately makes $D_1$ conductive and $C_2$ charges very rapidly so cutting off $D_2$ once again. During each succeeding field pulse the pattern is repeated. It will be realised that the rise of $D_2$ anode current follows the latter part of the discharge curve of $C_2$ (see shaded part of waveform (c)) and cannot be very sharp. Thus the shape of the leading edge of the output voltage pulse is determined by the time-constant of $C_2R_2$. The fall in anode current of $D_2$ is much sharper. This is because $C_2$ charges through the low resistance of the diode $D_1$ making the charging time constant of $C_2$ much faster than the discharging time-constant. The trailing edges of the output pulses are therefore much sharper.

If these negative-going output pulses are differentiated, steep positive-going pulses are produced on the trailing edges. Approximate waveforms for the circuit are shown. If the waveform at $D_1$ cathode is regarded as an input voltage applied to $D_2$ cathode, $D_2$ then takes on the appearance of a biased limiter which slices the waveform along a line whose level is fixed by the anode bias. The shaded tips of the cathode pulses show the periods of conduction of $D_2$. An output pulse is produced during the last part of each radiated field pulse. The circuit is capable of giving good quality interlace. It has the disadvantage of causing some degradation of the line pulses. Although the field oscillator is isolated by $D_2$ from the effects of these line pulses, the conduction of $D_1$ on each sync. pulse throws the capacitor $C_2$ in shunt with the sync. separator anode circuit. This results in the leading edges of line pulses being less sharp. Where flywheel synchronisation of the line timebase is employed, this fact does not cause any difficulty.

**Partial differentiator and triode clipper**

Fig. 13.9 shows an example of a circuit which makes use of a field pulse partial differentiator followed by a clipper. The differentiating network $C_1R_1$ is connected between the anode of the sync. separator and chassis in the normal way.

The triode clipper has a high value anode load resistor and is biased to cut-off by the cathode bias voltage developed by the pentode anode and screen currents. The cathode resistor is seen to be of 100 kΩ, by-passed by an 8 μF capacitor. The time constant of this network is $(8 \times 10^{-6} \times 100 \times 10^3) = 0{\cdot}8$ seconds so that, despite the fact that the pentode anode current is cut off for the most of the time, the cathode bias built up by its pulsed anode current produces a virtually steady cathode bias voltage.

The partially differentiated waveform across $R_1$ has the form described for 405- and 625-line signals respectively in Figs 11.9 and 11.10 of Chapter 11. As shown, during line pulses the greater part of the waveform lies below zero potential, extending only a very few volts positive to chassis on the trailing edges of the line pulses. The bias on the triode is such that these positive movements are too small to carry the net bias to a point less than cut-off, and there is therefore no output at the triode anode. During the field pulse chain the half-line pulses ride up above the zero voltage line to much higher positive potentials. Each one of these extends under the grid-base of the triode, as illustrated in Fig. 13.9(b). For each half-line pulse there is a pulse of anode current and a corresponding negative-going voltage pulse at the triode anode.

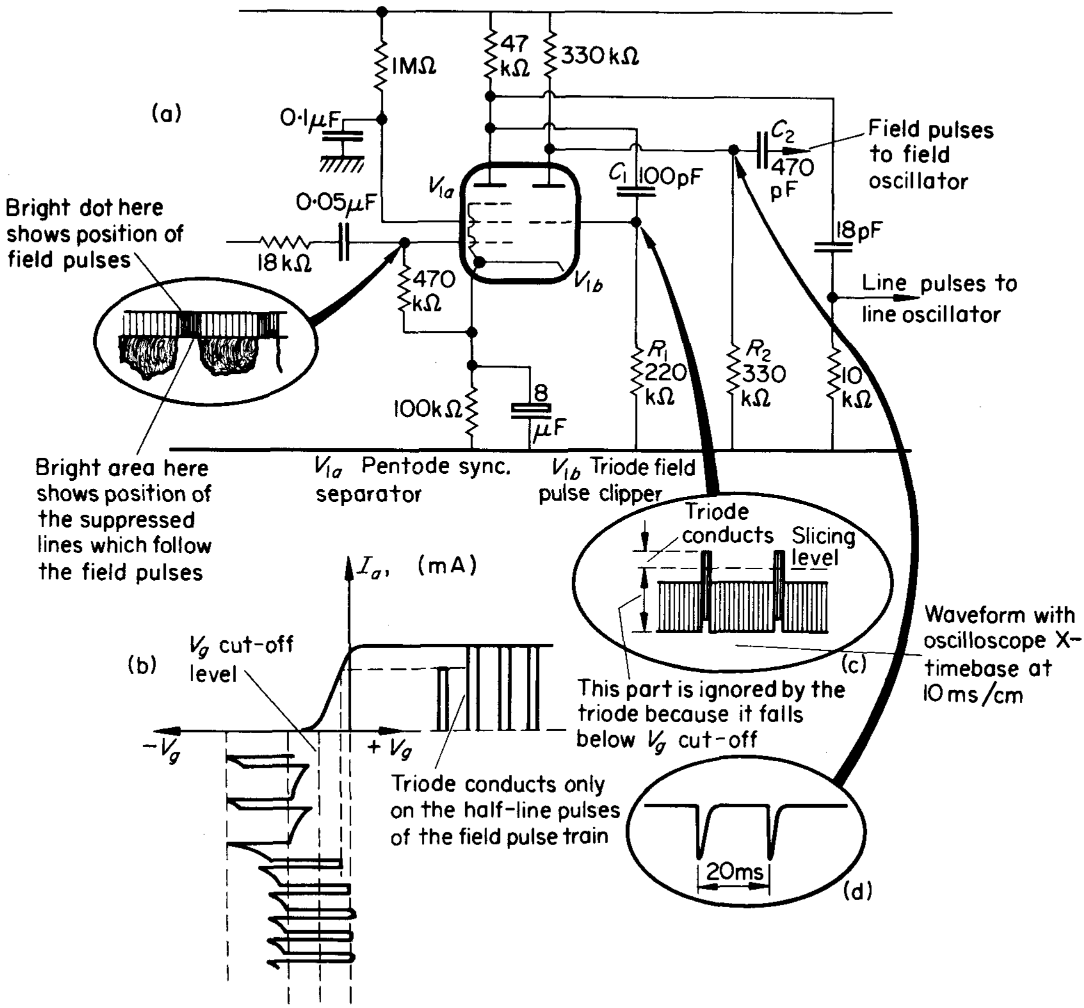



Fig. 13.9 Circuit employing a field pulse differentiating circuit ($C_1R_1$) followed by a triode field-pulse clipper. Fig. (b) shows how the triode performs the clipping action

$$\begin{aligned}\text{Time-constant of the field pulse differentiating circuit} &= CR \text{ seconds}\\ &= 100\times10^{-12}\times220\times10^{3}\times10^{6}\ \mu\text{s}\\ &= \underline{22\ \mu\text{s}}\end{aligned}$$

These output pulses are coupled by the d.c. blocking capacitor $C_2$ to the field oscillator circuit. The oscillator triggers on the first of the pulses.

The advantages and disadvantages of the field pulse partial differentiator have been discussed. It gives a very steep incisive pulse, with perfect timing on odd and even fields, but is not able to ride out interference pulses in the way that the slow-acting integrator can do. Where no equalising pulses are present it suffers the additional disadvantage of considerable differences in the trailing edges of the complete output waveform, which leads to the interlacing difficulties described.

**Field pulse processing in transistor circuits**

The same techniques are employed in transistor circuits; both to separate the field pulses from the combined line and field pulses, and to process the integrated waveform. The time constants of differentiating and integrating networks are of the same order, but to establish these in low impedance current-operated transistor circuitry it is necessary to use lower resistor values and correspondingly higher capacitor values.

*Integrator followed by diode clipper.* A typical transistor circuit is shown in Fig. 13.10(a). Here $R_1$ and $C_1$ form an integrator, connected across the output terminals of the transistor sync. separator TR1. The time constant of the integrator is $C_1 . R_1$ seconds $= 4{\cdot}7 \times 10^3 \times 4700 \times 10^{-12} \times 10^6$ μs $= 22$ μs. This is a reasonable value both for the 405-line and 625-line signals. The behaviour of integrators has been discussed fully in Chapter 11 and the only difference in transistor circuitry, apart from the lower ratio of resistance to capacitance, is one of voltage polarities.

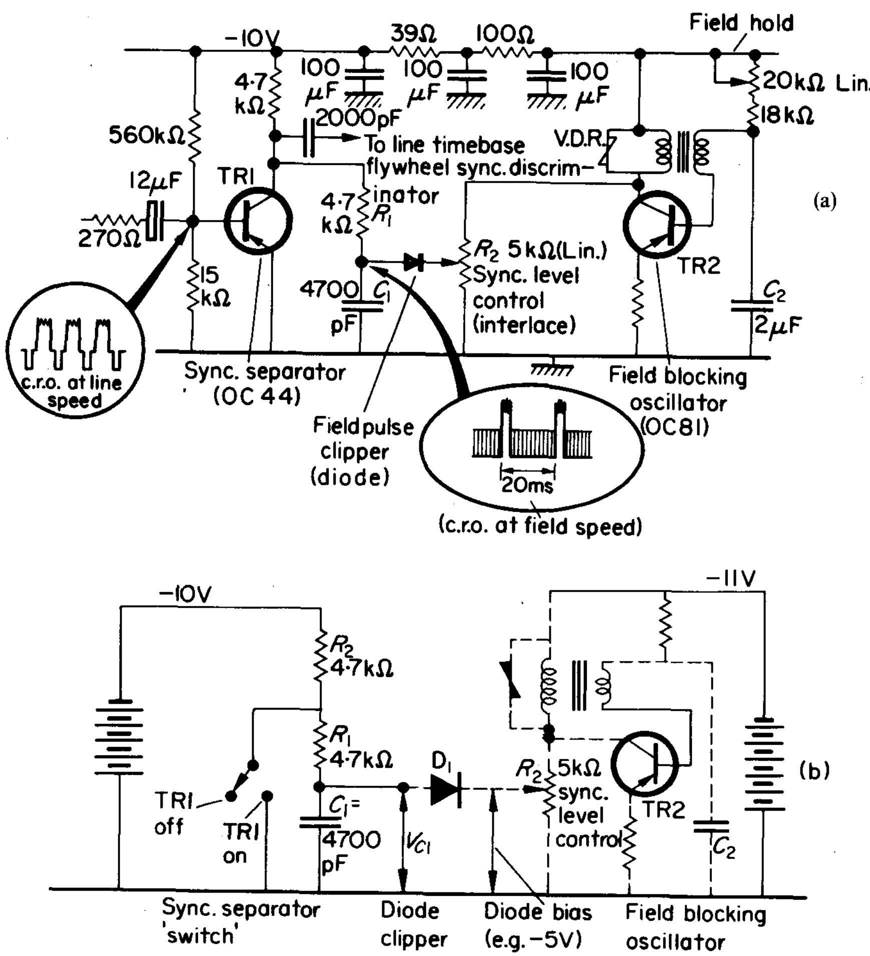



Fig. 13.10(a) Transistor sync. separator followed by field-pulse integrator and diode clipper
Fig. 13.10(b) Simplified representation of the circuit of Fig. 13.10(a)

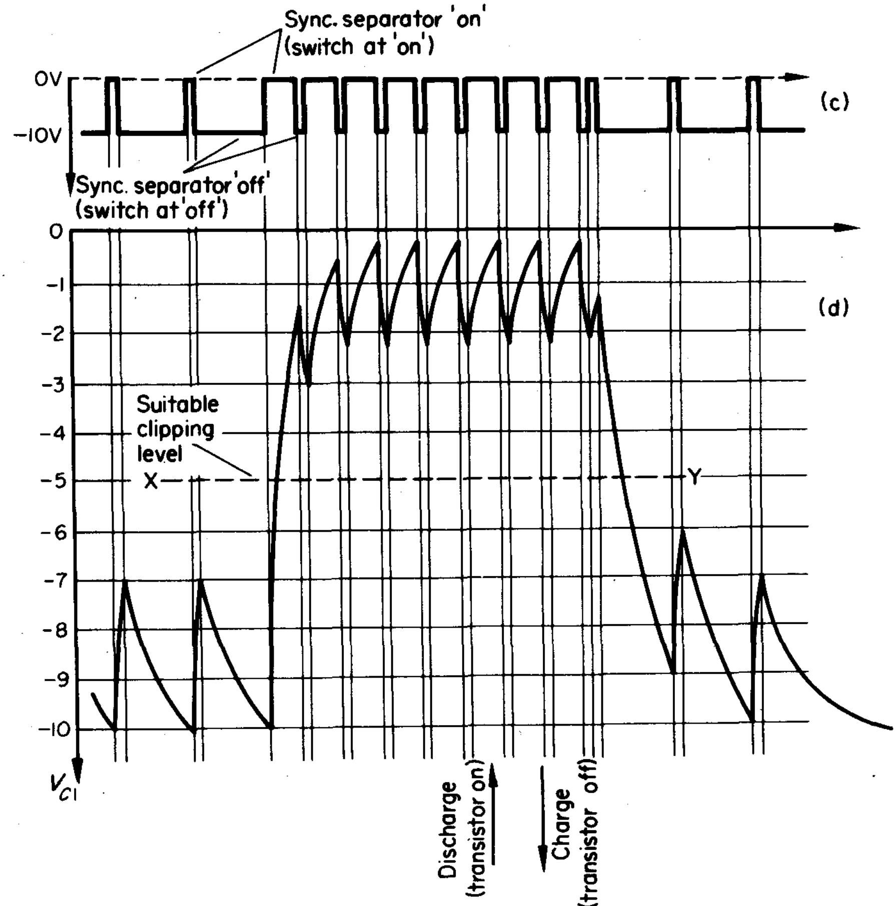



Fig. 13.10(c) Idealised waveform governing the charge and discharge of $C_1$

Fig. 13.10(d) Approximate shape of the composite waveform built up across $C_1$ in the simplified circuit of Fig. 13.10(b)

*Note.* $C_1$ charging time constant is $C_1(R_1+R_2) = 44\ \mu s$.
$C_1$ discharging time constant is $C_1R_1 = 22\ \mu s$.

Between sync. pulses, when the base of TR2 is held positive to emitter by the charge on the series feed capacitor, the only collector current is leakage current. The volts drop across the collector load resistor is therefore very small, which means that the voltage present on the collector is only a little less than the full negative supply line voltage $-V_{cc}$. During sync. pulses, however, when the base is fully driven by the negative going pulses, the collector current rises sharply to maximum and the transistor 'bottoms'; i.e. the collector voltage with respect to chassis is almost zero.

The situation facing the integrator may be represented by the simplified circuit shown in Fig. 13.10(b). The transistor is replaced by a switch which is open between sync. pulses and closed when they are present. When the switch is open $C_1$ charges through the two resistors $R_1$ and $R_2$ in series; the top plate of $C_1$ being negative to chassis. When the switch is closed $C_1$ discharges through $R_1$ alone. It is immediately apparent that the charging time-constant is

longer than the discharging time-constant of 22 μs. The discharge time-constant is $(R_1+R_2).C_1$ seconds, which is 44 μs since $R_1=R_2$, and $C_1R_1=22$ μs.

An approximate picture of the waveform built up across $C_1$ may be arrived at by the same technique used in Chapter 11. At the end of a normal picture line it may be assumed that $C_1$ is fully charged to the supply line voltage $-V_{cc}$; i.e. the top plate of $C_1$ rests at $-10$ V with respect to chassis. In Fig. 13.10(c) the 405-line field-pulse waveform is shown as it appears at the end of even fields. During each sync. pulse the voltage applied to $R_1$ and $C_1$ is zero, so that as $C_1$ discharges the potential of the top plate moves *upwards* in the positive direction, from $-10$ V towards 0 V. Between the pulses, $R_1$ is connected in series with $R_2$ to a potential of $-10$ V and the potential on the top plate of $C_1$ now moves *downwards* away from 0 V towards $-10$ V. It is necessary to perceive that as the voltage across the capacitor 'falls', the potential of the upper plate relative to chassis 'rises' in the positive-going direction from $-10$ V towards 0 V. The waveform produced across $C_1$ during the field pulse sequence is seen to be 'positive-going', even though the whole of it is disposed *below* chassis potential as an entirely *negative* waveform. In the corresponding valve circuit a negative-going composite field pulse waveform is produced by an integrator connected to the anode of the sync. separator but in this case the whole of it is disposed *above* chassis potential as an entirely *positive* voltage.

The step-by-step voltage changes which take place across $C_1$ may be deduced as demonstrated in Chapter 11. The formulae governing the 'downward' charging movements and 'upward' discharging movements are shown in the footnote below.*

Returning again to the circuit diagram, the next step is to clip the integrated waveform above the level reached during line sync. pulses. As a basis for discussion, suppose that it is decided to clip the waveform at the $-5$ V level, as shown by the line XY in Fig. 13.10(d). A simple diode clipper is used to achieve this. The diode's anode is connected to $C_1$ whilst its cathode is returned to the slider of a potentiometer connected between the collector of the field oscillator transistor TR2 and chassis. In the case postulated, the slider would be set to a point giving a potential of $-5$ V relative to chassis.

Under these circumstances the diode remains back-biased and of high resistance throughout

* *Charging formula* $$V_{c_1} = -\left\{E+(10-E)\left(1-e^{-\frac{t}{C_1(R_1+R_2)}}\right)\right\}$$

*Discharging formula* $$V_{c_1} = -E.e^{-\frac{t}{C_1R_1}}$$

where $V_{c_1}$ is the voltage on $C_1$ at the end of the given time interval of $t$ seconds.
$E$ is the voltage on $C_1$ at the beginning of the given time interval.
$C_1 = 4700\times10^{-12}$ farads, $R_1 = R_2 = 4{\cdot}7$ kΩ, and e $= 2{\cdot}71828$

For example, in Fig. 13.10(d), for the discharging stroke on the first 40 μs field pulse; $t=40$ μs and $E=-10$ V.

$$\therefore\ V_{c_1} = -10.e^{-\frac{40\times10^{-6}}{22\times10^{-6}}} \simeq -10.e^{-2} \simeq -10\times0{\cdot}14 \simeq -1{\cdot}4\text{ V}$$

For the charging stroke on the half-line pulse which follows, $t=10$ μs, $E=1{\cdot}4$ V, and $C_1(R_1+R_2)=44$ μs.

$$\begin{aligned}\therefore\ V_{c_1} &= -\left\{1{\cdot}4+(10-1{\cdot}4)\left(1-e^{-\frac{10}{44}}\right)\right\}\\ &= -\{1{\cdot}4+8{\cdot}6(1-e^{-0{\cdot}23})\}\\ &= -\{1{\cdot}4+8{\cdot}6(1-0{\cdot}8)\}\\ &= -3{\cdot}1\text{ V.}\end{aligned}$$

all normal picture lines. During the field sync. pulse sequence, however, when the top plate of $C_1$ moves sharply less negative to chassis, the anode of the diode is carried positive to the cathode and it becomes highly conductive. Electrons flow down through the top part of $R_2$, from cathode to anode of the diode, and so to the top plate of $C_1$. That this is the current direction may be seen if it is imagined that the anode of the diode is momentarily connected to chassis. The electron flow through the primary winding of the blocking oscillator transformer causes a back e.m.f. across it which seeks to oppose the build-up of current. The direction of this e.m.f. is such that the collector end of the primary is driven positive to the remote end.

To sum up, the positive-going integrated field sync. pulse waveform built up across $C_1$ causes the diode to conduct and this results in a positive-going voltage pulse being applied to the collector of the blocking oscillator transistor. It remains to establish how it is that this triggers the oscillator to initiate the flyback stroke.

The mode of operation of blocking oscillators will be studied in detail in due course. At present it is sufficient to state that the blocking oscillator transistor is cut off during the scanning stroke and conductive during flyback. The function of the sync. pulse must be to cause the transistor to start conducting again at the end of one field scanning stroke, to initiate the next burst of oscillation. Once started, so vigorous is the oscillation, that a heavy base-current flows causing a positive cut-off bias to develop immediately across the 'auto-bias' capacitor $C_2$. This holds the transistor cut-off again for another complete field scanning period.

It is obvious that a non-conductive p.n.p. transistor is not affected by a positive-going pulse to its collector. The transformer, however, serves as a coupling transformer between collector and base. Since the transformer windings are connected to give positive feedback, it follows that the positive-going field sync. pulse across the primary winding must give rise to a negative-going pulse at the base. This is in the right direction to drive the p.n.p. transistor into conduction and the blocking oscillator action is set off.

In the circuit of Fig. 13.10(a), the diode's bias-setting potentiometer is labelled *sync. level control*. The manufacturer's instruction is that this should be adjusted for optimum interlace. The way in which the quality of interlace is affected by the clipping level was discussed fully in Chapter 12.

*Integrator followed by a transistor clipper-inverter.* A further example of transistor sync. separator and field pulse processing circuitry is shown in Fig. 13.11. Grid modulation of the c.r.t. is used in this particular receiver and a positive-going video signal of some 50 V peak-to-peak amplitude is present at the collector of the last video amplifier stage. Only about 20% of this signal amplitude is taken off for driving the sync. separator. This division of the signal is achieved by employing two series resistors in the video amplifier collector load circuit.

As illustrated in the diagram, the video signal presented to the base of the sync. separator is of 10 V peak-to-peak amplitude, with the sync. pulses as the most negative part of the waveform. The sync. separator is fired into conduction in the normal way by the tips of the negative-going sync. pulses, and cut off for the remainder of the video signal by the positive-to-chassis back-bias of about 4·5 V which develops across the 0·25 μF series capacitor shown on the diagram.

If the transistor behaved as a perfect switch, the collector voltage would change abruptly on the arrival of each sync. pulse, and move from the supply voltage of −8·5 V up to the chassis potential of 0 V. The perfect square-edged waveform sketched in Fig. 13.10(c) would therefore be present at the collector. However, in the practical circuit, the combined effects of shunt capacitance across the collector circuit and the phenomenon of hole storage referred to in Chapter 10, modifies the waveform to produce pulses having slightly sloping leading edges and more pronounced slopes on the trailing edges. This may be seen clearly if the line pulses present

at the collector are viewed on the screen of an oscilloscope. The form of these pulses is sketched in the TR1 collector waveform 'balloon' on Fig. 13.11.

Field pulse separation is achieved by the integrator $R_3C_3$ which has a time-constant of $t = C_3R_3$ seconds $= 0{\cdot}01 \times 10^{-6} \times 8{\cdot}2 \times 10^3 \times 10^6$ μs $= 82$ μs. The charge on $C_3$ cannot change significantly during the brief interval of a line sync. pulse, as is illustrated on the diagram in the inset showing the integrator output waveform. During the successive field sync. pulses, the

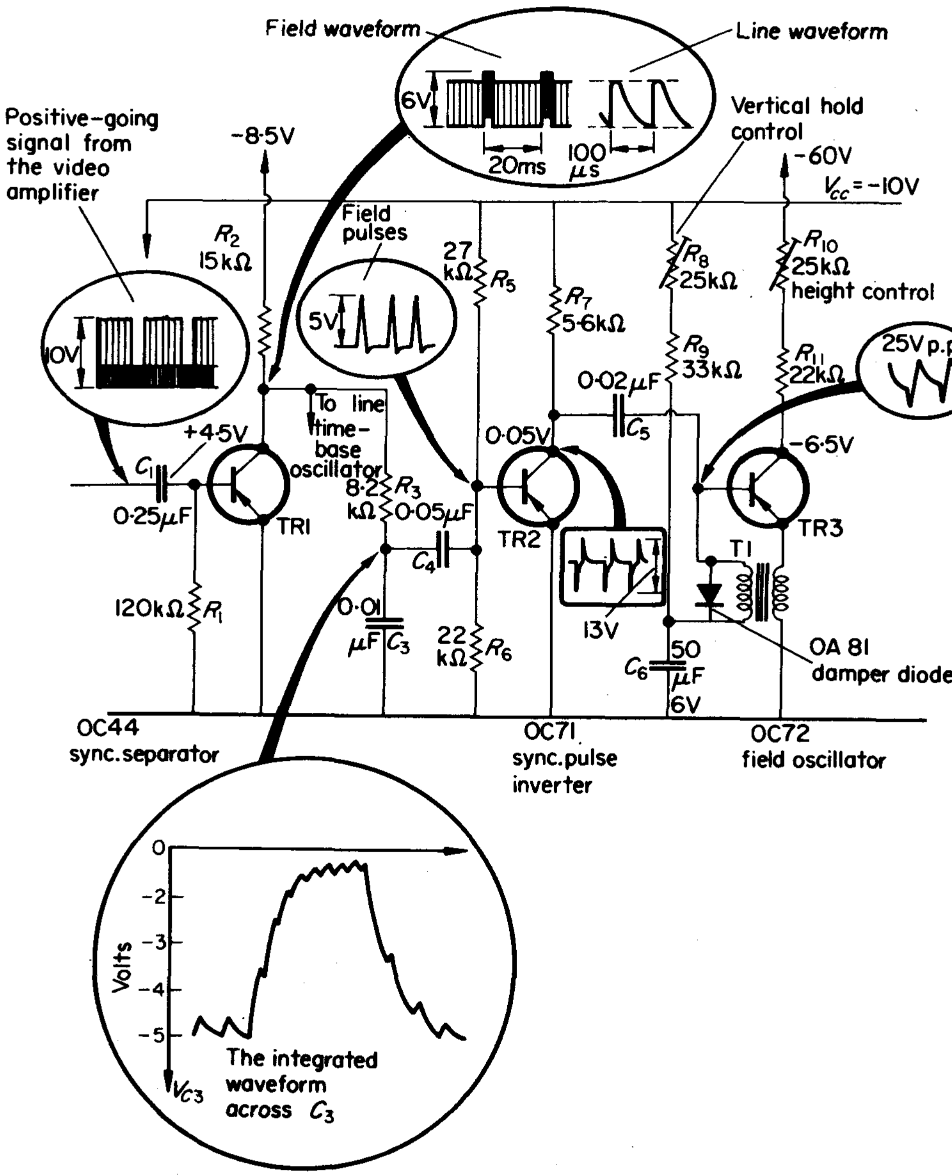



Fig. 13.11 Transistor circuit showing the sync. separator, field pulse integrator, field pulse clipper-invertor, and the *field timebase blocking oscillator*

voltage across $C_3$ moves upwards in a series of steps towards chassis potential. As in the previous circuit, the charging time-constant which controls the movement of voltage across $C_3$ when the sync. separator is cut off is longer than the discharging time constant, due to the inclusion of a second resistor $R_2$ of 15 kΩ in series with the integrator resistor $R_3$ of 8·2 kΩ. In this circuit the disparity is greater still and the charging time-constant is 230 µs. Because of this, the amount by which the voltage across $C_3$ 'slips backwards' negatively during the half-line pulses which separate the field pulses is only very small, and the resulting serrations in the composite field pulse are very shallow. When viewed on an oscilloscope with the X-timebase running at a low speed (e.g. 20 ms/cm), the composite waveform delivered by $C_4$ to the base of the field sync. pulse inverter TR2, shows up as a sharp positive-going pulse of some 5 V amplitude.

This second transistor is hard on during the inter-pulse period. As will be noted, there is no emitter resistor and the base is well forward biased by the potential divider $R_5R_6$. There is a 5·6 kΩ resistor in the collector lead which limits the maximum collector current to about 1·8 mA. In the inter-pulse periods, therefore, the transistor 'bottoms' and the collector voltage is almost zero.

The positive going movement of potential across $C_3$ during line pulses is so small that the change of potential transmitted to the base of TR2 by $C_4$ is insufficient to make any difference at all to the state of conduction of the transistor. The composite pulse formed by integrating the field sync. pulses, however, drives the base sharply positive to emitter and this causes the collector current to cut off suddenly. In turn this results in a sharp negative-going change of potential at the collector which moves from 0 V almost to the full supply line potential of −10 V.

The arrival, via the coupling capacitor $C_5$, of this sharp negative-going pulse at the base of the field blocking oscillator transistor TR3 triggers it into conduction, and the blocking oscillator fires immediately. The action of the blocking oscillator is such that the high amplitude first half-cycle of oscillation causes heavy base current, which charges up the coupling capacitor $C_5$ to produce at TR3 base a positive cut-off bias of some 6·8 V.

It is this which accounts for the positive-going trailing edge to the waveform shown at the collector of the field pulse inverter. When following the progress of synchronising pulses through a circuit with an oscilloscope, it must be remembered that at the point of application of the pulse to the oscillator, the waveform seen is a combination of the applied sync. pulse and the oscillatory waveform present at that point. For instructional or fault-finding purposes, it is sometimes a useful exercise to put the field timebase out of action in order to follow through the actual sync. pulse to its destination.

A point which should be noted before leaving Fig. 13.11 is that the transistor labelled sync. pulse inverter not only performs this function of producing a negative-going pulse from a positive-going one but also serves as a clipper, since it prevents line pulse information from reaching the field oscillator.

*Integrator d.c. coupled to the field oscillator.* Finally, Fig. 13.12 shows the video amplifier, sync. separator and field blocking oscillator circuitry of another transistor receiver. It is instructive to follow the path of the signal through the circuit by reference to the waveforms shown. Those labelled (a), (b), (c), (d) and (e) are of the form which should be seen on an oscilloscope with the X-timebase running at a slow enough speed to allow complete fields to be seen. As indicated before, a suitable speed for this purpose would be say 20 ms/cm. The additional sketches, in the balloons, give further information. At (c′) the waveform present at the sync. separator collector is examined with the X-timebase speeded up to allow individual line pulses to be observed. A convenient timebase setting here would be 50 µs/cm. This waveform should

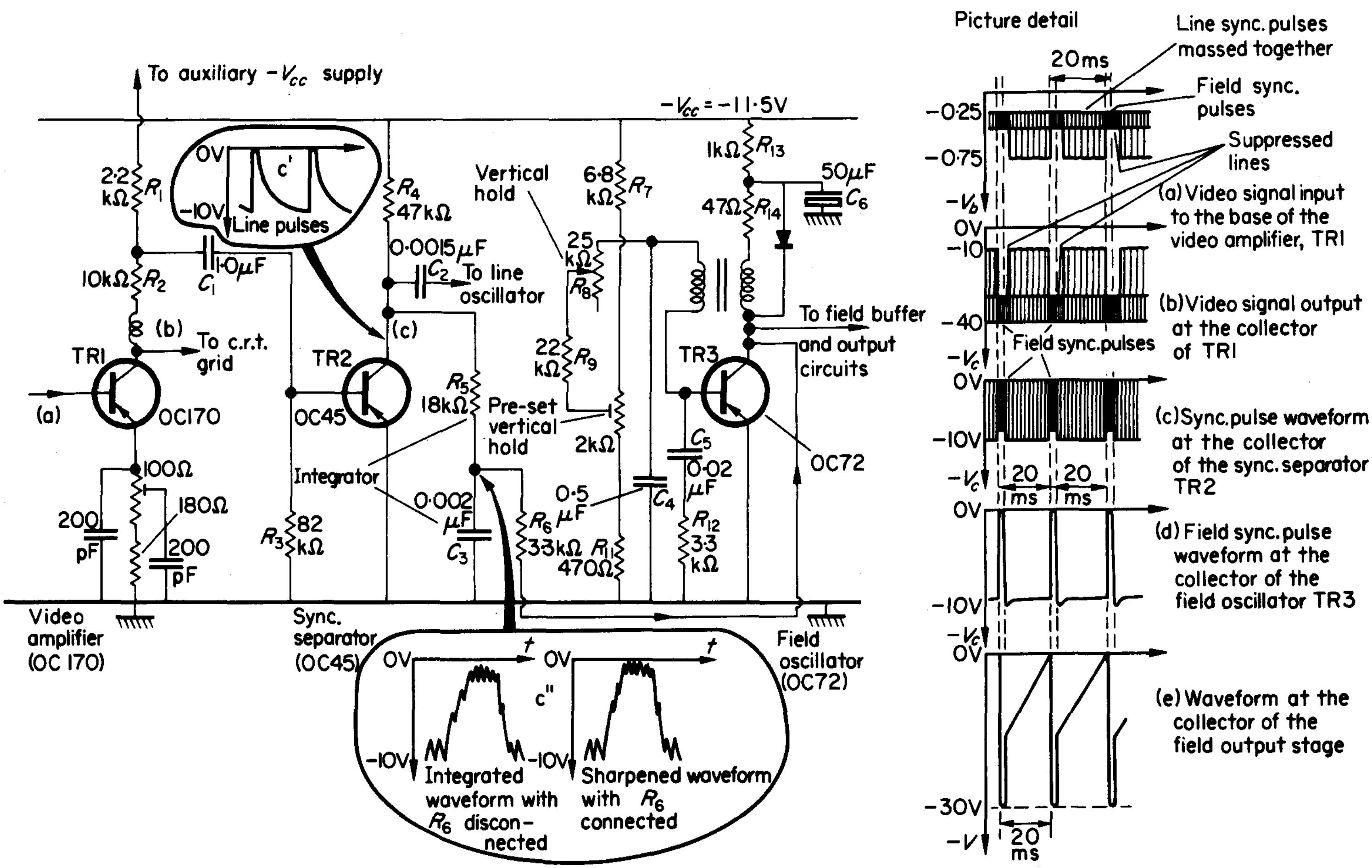



Fig. 13.12 Showing the progress of the video signal and the development of the field synchronising pulse in a transistor circuit, in which the field pulse integrator is d.c. coupled to the field blocking oscillator

(a) Video signal input to the base of the video amplifier TR1.
(b) Video signal output at the collector of TR1.
(c) Sync. pulse waveform at the collector of the sync. separator TR2.
(d) Field sync. pulse waveform at the collector of the field oscillator TR3.
(e) Waveform at the collector of the field output stage.

be contrasted with (*c*), which shows the corresponding waveform seen with the X-timebase running at 20 ms/cm. At (*c*) the line pulses are seen massed together between the field sync. pulse regions. During these latter periods, when line pulses are occurring at twice the normal line frequency, the oscilloscope trace is brighter. If, whilst looking at (*c*), the X-gain control is advanced, the waveform is expanded out and the individual field pulses may be seen. It is also instructive to do the same thing with the waveform (*b*) seen at the collector of the video amplifier. With the waveform closed up as at (*b*), the regions of the field pulses and post-field pulse suppressed lines may be discerned because the oscilloscope trace brightens up at the appropriate amplitude levels.

Returning to a description of the circuit itself, once again grid modulation of the c.r.t. is used and so a positive-going video signal is present at the collector of the last stage of the video amplifier. The no-signal voltages at the base and collector of this particular transistor are of the order $-0{\cdot}25$ V and $-40$ V, respectively, relative to chassis. On a normal picture the base voltage is driven negatively down to about $-0{\cdot}75$ V on peak whites. The increase of collector current causes a corresponding 'rise' in collector voltage from $-40$ V to the order of $-10$ V.

The negative-going sync. pulses delivered to the base of the sync. separator transistor TR2, cause the collector voltage to rise sharply to almost 0 V; i.e. the transistor bottoms. In the inter-pulse periods the transistor is cut off due to the charge on $C_1$ which holds the base of TR2 positive to emitter. This allows the collector voltage to return negatively, to a level almost equal to the supply line voltage of $-11{\cdot}5$ V. The inset at ($c'$) shows how the combined effects of shunt capacitance and hole storage cause the trailing edges of the pulses to decay comparatively slowly. However, as pointed out in Chapter 10, the switch-on time of a transistor is much faster than the switch-off time, and the all-important leading edges of the sync. pulses have the steep-fronted form required for efficient synchronisation of the timebases.

A field pulse integrator $R_5C_3$ is connected to the collector of the sync. separator. This has a time constant of $0{\cdot}002\times10^{-6}\times18\times10^{3}\times10^{6}$ $\mu$s$=36$ $\mu$s. The circuit differs from the two previous ones in an important respect. It will be noticed that the integrator capacitor $C_3$ is here d.c. coupled by the $3{\cdot}3$ k$\Omega$ resistor $R_6$ to the collector of the field blocking oscillator transistor TR3. The introduction of d.c. coupling between the integrator and the oscillator makes a difference to the mode of operation of the integrator.

The way in which the circuit operates is best studied by starting with the moment of arrival of the first field sync. pulse. At this moment the field blocking oscillator is still cut-off, but the scanning stroke is nearly at an end and TR3 is reaching a critical state just before the time it is due to be triggered into oscillation. The positive-going build up of the composite field pulse produced by the switching action of the sync. separator transistor TR2, goes ahead in the normal way. The charging time-constant, which affects the 'negative slide-back' of potential across $C_3$ during the half-line pulses, is, however, modified somewhat by the introduction of a second d.c. path from the top plate of $C_3$ to the $-V_{cc}$ supply line. This second d.c. circuit is via $R_6$, the blocking oscillator transformer primary, $R_{14}$ and $R_{13}$. The three resistors are of low value compared with the TR2 collector resistor $R_4$ which is of 47 k$\Omega$. At first sight it might appear that the second d.c. path would cause a marked reduction in the charging time constant of $C_3$, but in fact, the inductance of the transformer primary is high and takes charge of the rate at which current may flow via this path to $C_3$. If this were not so it is obvious that there would be no build up of a composite pulse across $C_3$, since $C_3$ would gain during the half-line pulses, the charge lost during the field sync. pulses.

The positive-going composite pulse built up across $C_3$ therefore causes a back-e.m.f. across the transformer primary and this in turn gives rise to a negative-going voltage across the

secondary winding. The base of TR3 is driven sharply negative to the emitter and the blocking oscillator action is triggered off. The result of this is that once TR3 'fires' and becomes highly conductive, $R_6$ is effectively strapped in shunt with $C_3$ since the collector of TR3 is virtually at chassis potential. It follows that once the composite field pulse across $C_3$ reaches the level which triggers the blocking oscillator action, the discharging time-constant of $C_3$ is reduced from 36 μs to a value of the order of 6 μs. If the pulse waveform across $C_3$ is viewed under working conditions therefore, a positive-going pulse of very fast rise time is seen, and the voltage on the top plate of $C_3$ moves more sharply from the supply line level of $-10$ V up to the chassis potential of 0 V than it would do if $R_6$ were disconnected.

The last of the column of waveforms in Fig. 13.12, marked (*e*), shows the voltage present at the collector of the field output stage. The scanning stroke shows up as a linear change of voltage, and the way in which the flyback stroke coincides with the trigger pulse which initiates it, may be seen by comparing (*d*) with (*e*).

# Appendix

### Equivalent circuits

For those not familiar with equivalent circuits, Fig. 6.3(b) may be arrived at by the following logical steps:

(a) Since the H.T. line is decoupled to the chassis by a capacitor offering a very low reactance to signal frequencies, the H.T. line and the chassis may be considered to be strapped together, or common, as far as a.c. is concerned. Any component connected to the H.T. line may thus be represented as connected to chassis. The H.T. line of the circuit in Fig. A.1(a) may therefore be imagined to be folded down and 'pinned' to the chassis, which then brings $R_l$ in shunt with the valve. This is illustrated in Fig. A.1(b).

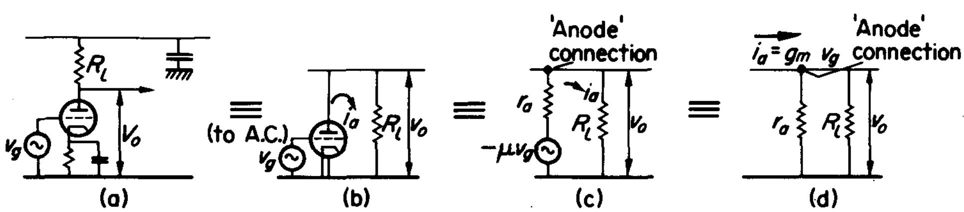



Fig. A.1

(b) But, if a voltage $v_g$ is applied between grid and cathode of a valve having an amplification factor of $\mu$, then the valve develops an output of $\mu.v_g$ volts. This is often written as $-\mu v_g$ to take account of the signal inversion between grid and anode. Since the resistance offered by a valve to alternating current is equal to its a.c. slope resistance $r_a$, it follows that the valve may be represented as a generator of internal resistance $r_a$ ohms, developing an output of $\mu.v_g$ volts. This leads directly to Fig. A.1(c) and to Fig. 6.3(b).

(c) Often, particularly with pentode valves, an alternative 'constant current' form of equivalent circuit is more convenient to use. This is shown in Fig. A.1(d) and Fig. 6.3(c). It may be arrived at very easily from the original equivalent circuit (Fig. A.1(c)) as follows:

The alternating current $i_a$ set up by the generator of voltage $u.v_g$ is clearly given by:

$$i_a = \frac{\mu.v_g}{r_a+R_1}$$

The output voltage $V_o$ is then:

$$v_o = i_aR_l = \frac{\mu.v_g}{r_a+R_l}\times R_l$$

Since $\mu = g_m r_a$ this may be rewritten in the form:

$$v_o = \frac{g_m r_a v_g}{r_a + R_l} . R_l = g_m v_g \times \frac{r_a R_l}{r_a + R_l}$$

A study of this last expression shows that the output voltage is the same as that which would be set up by a current of $g_m v_g$ amps flowing into a parallel combination of two resistors having values of $r_a$ and $R_l$ ohms. This leads to the equivalent circuit shown in Fig. A.1(d) and Fig. 6.3(c).

Obviously, any components connected between anode and chassis, such as the series combination $C_c$ and $R_g$, in Fig. 6.3(a) are shown in shunt with $R_l$ in the equivalent circuits.

For the benefit of those interested, some of the formulae used in Chapter 8 are now derived.

**Relative gain at H.F.**

*Object. To express the gain at H.F. as a ratio against the gain at M.F.*

*Step 1.* Find an expression for the gain at M.F.
This is shown in the text on p. 94 to be given by:

$$\text{Gain at M.F.} = g_m . R_t \quad \ldots \quad (1)$$

*Step 2.* Find an expression for the gain at H.F.

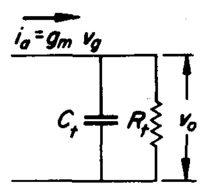



Fig. A.2

From Fig. 6.5 and Fig. A.2 it is clear that:

$$v_o = i_a Z_t = g_m v_g . Z_t$$

$$\frac{v_o}{v_g} = \text{Gain at H.F.} = g_m Z_t$$

But

$$Z_t = \frac{R_t\left(\frac{1}{j\omega C_t}\right)}{R_t + \frac{1}{j\omega C_t}} = \frac{R_t}{j\omega C_t R_t + 1}$$

$$\therefore \text{Gain at H.F.} = g_m Z_t = \frac{g_m R_t}{1 + j\omega C_t R_t} = \frac{g_m R_t}{\sqrt{1 + (\omega C_t R_t)^2}}$$

But, from (1),

$$g_m R_t = \text{Gain at M.F.}$$

$$\therefore \text{Gain at H.F.} = \frac{\text{Gain at M.F.}}{\sqrt{1 + (\omega C_t R_t)^2}} \quad \ldots \quad (2)$$

*Object. To find the frequency $f_2$ at which the gain at H.F. falls by 3 dB on the gain at M.F.* (see $f_2$ on Fig. 6.3(d))

This occurs when the ratio $\dfrac{\text{Gain at H.F.}}{\text{Gain at M.F.}} = \dfrac{1}{\sqrt{2}}$. Examination of Eq. (2) shows that this is so when:

$$\frac{1}{\sqrt{1+(\omega C_t R_t)^2}} = \frac{1}{\sqrt{2}}$$

i.e. when $$\omega C_t R_t = 1$$

Thus $$2\pi f_2 C_t R_t = 1$$

$$f_2 = \frac{1}{2\pi C_t R_t} \qquad (3)$$

*Object. To study the effect when $R_l$ is of low value*

Now, $R_t$ is the resistance of $r_a$, $R_l$ and $R_g$ in parallel.
Thus, if $R_l$ is very low, cf. $r_a$ and $R_g$, then $R_t \simeq R_l$.
In this case, $R_l$ may be substituted for $R_t$ in Eqs. (1), (2) and (3).

$$\therefore \text{ Gain at M.F.} = g_m R_l \qquad (1')$$

$$\text{Gain at H.F.} = \frac{\text{Gain at M.F.}}{\sqrt{1+(\omega C_t R_l)^2}} \qquad (2')$$

$$f_2 = \frac{1}{2\pi C_t R_l} \qquad (3')$$

**Relative gain at L.F.**

*Object. To express the gain at L.F. as a ratio against the gain at M.F.*

The 'constant voltage' form of the equivalent circuit is more convenient for this case. Fig. 6.6 and Fig. A.3(a) show the effective circuit.

Thévenin's Theorem may usefully be applied here. This states that:

The current which passes through a load impedance connected to two terminals A and B of a network of generators and impedances is the same as if the load were connected to a simple constant-voltage generator whose e.m.f. is the open-circuit voltage measured across A and B with the load disconnected, and whose internal impedance is the impedance seen looking back into the open circuited terminals with all generators replaced by impedances equal to their own internal impedances.

*Step 1.* To find the open circuit voltage $v'$ at AB (see Fig. A.3(b)).

This is simply the voltage across $R_l$ which would result from the current driven through $r_a$ and $R_l$ in series by the voltage $\mu v_g$.

Thus $$v' = \frac{\mu v_g}{r_a + R_l} \times R_l \qquad (4)$$

*Step 2.* To find the open circuit impedance at AB with generators replaced by their internal impedances (see Fig. A.3(c)).

From the diagram, it is clear that:

$$Z' = \frac{r_a R_l}{r_a + R_l} + \frac{1}{j\omega C_c} \quad . \quad . \quad . \quad . \quad . \quad . \quad (5)$$

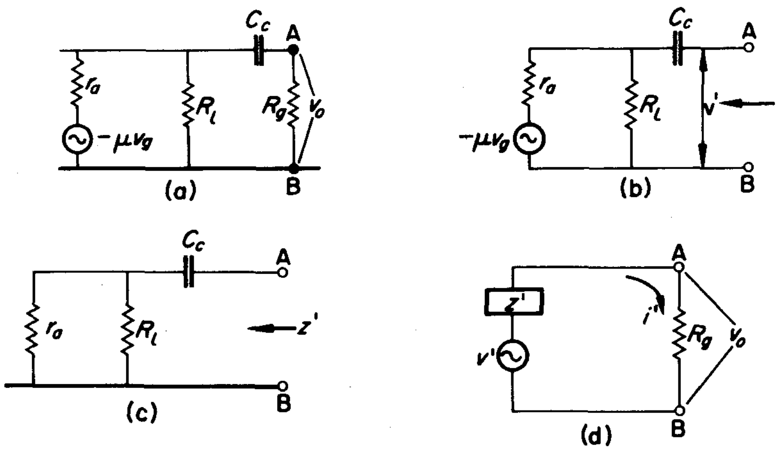



Fig. A.3

*Step 3.* To find the current $i'$ in $R_g$ due to a generator of e.m.f. given by Eq. (4) whose impedance is that given by Eq. (5).

$$i' = \frac{v'}{Z' + R_g} = \frac{\dfrac{\mu v_g R_l}{r_a + R_l}}{\dfrac{r_a R_l}{r_a + R_l} + \dfrac{1}{j\omega C_c} + R_g} \quad . \quad . \quad . \quad . \quad . \quad (6)$$

*Step 4.* To find $v_o$, and hence the ratio $v_o/v_g$, which is the required gain at L.F.

From Fig. A.3(d) it is seen that:

$$v_o = i' R_g = \frac{\dfrac{\mu v_g R_l}{r_a + R_l}}{\dfrac{r_a R_l}{r_a + R_l} + R_g + \dfrac{1}{j\omega C_c}} \times R_g$$

$$\therefore \frac{v_o}{v_g} = \text{Gain at L.F.} = \frac{\dfrac{\mu R_l R_g}{r_a + R_l}}{\dfrac{r_a R_l}{r_a + R_l} + R_g + \dfrac{1}{j\omega C_c}} \quad . \quad . \quad . \quad . \quad (7)$$

*Step 5.* To simplify Eq. (6) and express the ratio of gain at L.F. to gain at M.F. in a convenient form.

Substitute $R'$ for $\left(\frac{r_a R_l}{r_a+R_l}+R_g\right)$

$$\therefore \text{ Gain at L.F.} = \frac{\frac{\mu R_l R_g}{r_a+R_l}}{R'+\frac{1}{j\omega C_c}}$$

$$\text{Gain at L.F.} = \frac{\frac{\mu R_l R_g}{r_a+R_l}}{R'\left(1+\frac{1}{j\omega C_c R'}\right)}$$

$$= \frac{\mu R_l R_g}{(r_a+R_l)R'\left(1+\frac{1}{j\omega C_c R'}\right)}$$

$$= \frac{g_m r_a R_l R_g}{(r_a+R_l)\left\{\frac{r_a R_l}{r_a+R_l}+R_g\right\}\left\{1+\frac{1}{j\omega C_c R'}\right\}} \quad \text{because } \mu = g_m r_a$$

$$= \frac{g_m r_a R_l R_g}{r_a R_l + r_a R_g + R_g R_l\left(1+\frac{1}{j\omega C_c R'}\right)}$$

But
$$\frac{r_a R_l R_g}{r_a R_l + r_a R_g + R_g R_l} = R_t$$

where $R_t$ is the parallel resistance of $r_a$, $R_l$ and $R_g$

$$\therefore \text{ Gain at L.F.} = \frac{g_m R_t}{\sqrt{1+\left(\frac{1}{\omega C_c R'}\right)^2}} = \frac{\text{Gain at M.F.}}{\sqrt{1+\left(\frac{1}{\omega C_c R'}\right)^2}} \quad . \quad . \quad (8)$$

where
$$R' = \left\{\frac{r_a R_l}{r_a+R_l}+R_g\right\}$$

*Object. To find the frequency $f_1$ at which the gain at L.F. falls by 3 dB on the gain at M.F.* (see $f_1$ in Fig. 6.3(d))

$$\text{This occurs when } \frac{\text{Gain at L.F.}}{\text{Gain at M.F.}} = \frac{1}{\sqrt{2}}$$

Examination of Eq. (8) shows that this is so when:

$$\frac{1}{\sqrt{1+\left(\frac{1}{\omega C_c R'}\right)^2}} = \frac{1}{\sqrt{2}}, \quad \text{i.e. when} \quad \frac{1}{\omega C_c R'} = 1$$

Thus $$\frac{1}{2\pi f_1 C_c R'} = 1 \quad \text{and} \quad f_1 = \frac{1}{2\pi C_c R'} \qquad (9)$$

*Object. To study the effect when $R_l$ is of low value*

If $R_l$ is low, then $$R' = \left\{\frac{r_a R_l}{r_a + R_l} + R_g\right\} \simeq R_g$$

Thus (8) becomes: $$\text{Gain at L.F.} = \frac{\text{Gain at M.F.}}{\sqrt{1+\left(\frac{1}{\omega C_c R_g}\right)^2}} \qquad (8')$$

and (9) becomes:

$$f_1 = \frac{1}{2\pi C_c R_g} \qquad (9')$$

*Simplified analysis at L.F.*

If $R_l$ is known to be very low compared with $r_a$ and $R_g$, then the analysis for the gain at L.F. may be much simplified as follows:

The resistance of $R_l$ is so much lower than the impedance of the arm containing $C_c$ and $R_g$ that the current $i_a = g_m v_g$ may be assumed to flow entirely through $R_l$ (see Fig. A.4). This sets

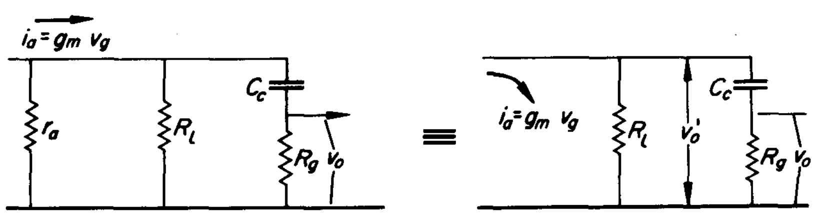



Fig. A.4

up a voltage $v_o'$ across $R_l$, and $C_c$ and $R_g$ may be treated as a potential divider connected across this voltage $v_o'$.

Thus $$v_o' = i_a R_l = g_m v_g R_l$$

but $$v_o = \frac{v_o'}{\sqrt{R_g^2 + \left(\frac{1}{\omega C_c}\right)^2}} \times R_g = \frac{g_m v_g R_l R_g}{\sqrt{R_g^2 + \left(\frac{1}{\omega C_c}\right)^2}}$$

$$\therefore \frac{v_o}{v_g} = \text{Gain at L.F.} = \frac{g_m R_l R_g}{\sqrt{R_g^2 + \left(\frac{1}{\omega C_c}\right)^2}}$$

Dividing by $R_g$ yields:

$$\text{Gain at L.F.} = \frac{g_m R_l}{\sqrt{1+\left(\frac{1}{\omega C_c R_g}\right)^2}}$$

But, when $R_l$ is of low value, $g_m R_l$=Gain at M.F.

$$\therefore \text{ Gain at L.F.} = \frac{\text{Gain at M.F.}}{\sqrt{1+\left(\frac{1}{\omega C_c R_g}\right)^2}} \quad \text{(see (8'))}$$

and $f_1 = 1/(2\pi C_c R_g)$ is the frequency at which the denominator $=\sqrt{2}$, and the gain falls by 3 dB.

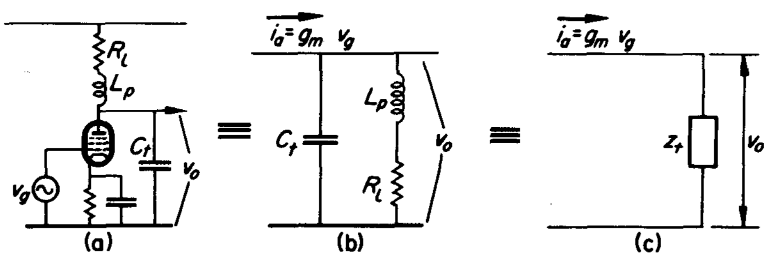
**The shunt peaking coil**

*Object. To obtain an expression for the ratio of the gain at H.F. to the gain at M.F. when a shunt peaking coil is included in the anode circuit*

*Step 1.* To examine the design formula $L_p = nR_l^2 C_t$.

(a) As is seen in Fig. 6.10 and Fig. A.5 the anode circuit takes on the *appearance* of a parallel tuned circuit. Such a circuit resonates at a frequency given by the formula:

$$f_r = \frac{1}{2\pi}\sqrt{\left(\frac{1}{L_p C_t} - \frac{R_l^2}{L_p^2}\right)} \quad . \quad . \quad . \quad . \quad . \quad . \quad (10)$$




Basic circuit

Equivalent circuit when $R_l \ll r_a$

Fig. A.5

(b) But, as stated in the text, the selective properties of a tuned circuit are *not* required. Suppose, therefore, that as a starting point the circuit is made *non-resonant* by making

$$\left(\frac{1}{L_p C_t} - \frac{R_l^2}{L_p^2}\right) = 0 \quad \text{(see (10))}$$

For a given value of $R_l$ and $C_t$, the inductance needed to establish this condition is given by:

$$\frac{1}{L_p C_t} = \frac{R_l^2}{L_p^2}$$

from which

$$L_p = R_l^2 C_t \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad . \quad (11)$$

(c) A study of Eq. (10) shows that if $L_p$ is made greater than the value shown in Eq. (11), the bracket under the root sign is positive and the circuit *does* resonate; but if $L_p$ is smaller, the bracket is negative and the circuit is non-resonant. By introducing into Eq. (11) a numerical

factor $n$, where $n$ may be 'less than', 'equal to' or 'greater than' unity, any required result may be achieved. This leads to the expression

$$L_p = nR_l^2C_t \quad . \quad . \quad . \quad . \quad . \quad . \quad . \qquad (12)$$

(d) The result may be summarized by saying that the anode circuit only behaves as a parallel tuned circuit if the factor $n$ is greater than unity.

In most practical circuits $n$ has a value less than unity, from which it follows that it is incorrect to attribute (as is sometimes done) the beneficial effect of a shunt peaking coil to the properties of a parallel resonant circuit formed by $L_p$, $C_t$ and $R_l$. This misconception arises partly because the response curves for a video amplifier employing a shunt peaking coil tend to show humps at the upper end of the frequency range, and these suggest that the anode circuit is resonating. Reference to Fig. 6.11 shows, however, that there is a small hump even when the factor $n=0{\cdot}5$, and quite a pronounced one when $n=1$. At both these values of $n$, the anode circuit is non-resonant and the humps must not be regarded as evidence that the amplifier is behaving as a tuned anode amplifier in this region. When $n$ is slightly greater than unity the circuit *is* resonant, but the corresponding resonant frequency is very low and certainly does not coincide with the hump at the H.F. end of the response curve.

*Step 2.* To arrive at an expression for the gain at H.F.

$$v_o = i_aZ_t = g_mv_gZ_t$$

$$\text{Gain at H.F.} = \frac{v_o}{v_g} = g_mZ_t \quad . \quad . \quad . \quad . \quad . \quad . \qquad (13)$$

But

$$Z_t = \frac{(R_l+j\omega L_p)\dfrac{1}{j\omega C_t}}{R_l+j\omega L_p+\dfrac{1}{j\omega C_t}}$$

$$\therefore\ Z_t = \frac{(R_l+j\omega L_p)}{j\omega C_tR_l+1-\omega^2L_pC_t} \quad \text{(multiplying top and bottom by } j\omega C_t\text{)}$$

Thus

$$\text{Gain at H.F.} = g_mZ_t = \frac{g_m(R_l+j\omega L_p)}{1-\omega^2L_pC_t+j\omega C_tR_l}$$

$$= \frac{g_mR_l\left(1+\dfrac{j\omega L_p}{R_l}\right)}{1-\omega^2L_pC_t+j\omega C_tR_l} \quad . \quad . \quad . \quad . \qquad (14)$$

*Step 3.* To arrive at an expression for the ratio of the gain at H.F. to the gain at M.F. As established, we have that:

$$\text{Gain at M.F.} = g_mR_l$$

$$\therefore\ \frac{\text{Gain at H.F.}}{\text{Gain at M.F.}} = \frac{\left(1+\dfrac{j\omega L_p}{R_l}\right)}{1-\omega^2L_pC_t+j\omega C_tR_l} \quad . \quad . \quad . \quad . \qquad (15)$$

This expression may be more conveniently expressed in terms of $n$ and the ratio $\omega/\omega_2$, where $n=L_p/(R_l^2C_t)$ (from (12)) and $\omega_2=1/(R_lC_t)$ (from (3′)).

Thus, substituting $nR_l^2C_t$ for $L_p$ in Eq. (15) yields:

$$\frac{\text{Gain at H.F.}}{\text{Gain at M.F.}}=\frac{1+j\omega n\,.\,R_lC_t}{1-\omega^2nR_l^2C_t^2+j\omega C_tR_l}$$

Also, substituting $1/\omega_2$ for $R_lC_t$ yields:

$$\frac{\text{Gain at H.F.}}{\text{Gain at M.F.}}=\frac{1+jn\left(\dfrac{\omega}{\omega_2}\right)}{1-n\left(\dfrac{\omega}{\omega_2}\right)^2+j\left(\dfrac{\omega}{\omega_2}\right)} \quad . \quad . \quad . \quad . \quad (16)$$

Rationalising then gives:

$$\frac{\text{Gain at H.F.}}{\text{Gain at M.F.}}=\frac{\sqrt{1+n^2\left(\dfrac{\omega}{\omega_2}\right)^2}}{\sqrt{\left\{1-n\left(\dfrac{\omega}{\omega_2}\right)^2\right\}^2+\left(\dfrac{\omega}{\omega_2}\right)^2}} \quad . \quad . \quad . \quad . \quad (17)$$

The response curves in the text show the relative gain in decibels. Since the expressions above refer to voltage gain, the voltage gain ratio shown as Eq. (17) may be expressed in decibels as follows:

$$\text{Relative gain in decibels} = 20\log_{10}\frac{\text{Gain at H.F.}}{\text{Gain at M.F.}}$$

$$=20\log_{10}\frac{\sqrt{1+n^2\left(\dfrac{\omega}{\omega_2}\right)^2}}{\sqrt{\left\{1-n\left(\dfrac{\omega}{\omega_2}\right)^2\right\}^2+\left(\dfrac{\omega}{\omega_2}\right)^2}}$$

$$=10\log_{10}\frac{1+n^2\left(\dfrac{\omega}{\omega_2}\right)^2}{\left\{1-n\left(\dfrac{\omega}{\omega_2}\right)^2\right\}^2+\left(\dfrac{\omega}{\omega_2}\right)^2} \quad . \quad (18)$$

The curves of Fig. 6.11 show approximate plots of this expression for various fixed values of $n$. It will be observed that the ratio $\omega/\omega_2=f/f_2$ is the ratio of the test frequency $f$ to the frequency $f_2$ at which the gain of the uncompensated amplifier falls by 3 dB. This device makes the curves universally applicable to any given amplifier.

Thus, taking a specific case, in a British 625-line receiver video amplifier, $f_2$ may be say 5 Mc/s, and the $x$-axis may be directly calibrated in frequency by substituting this value of $f_2$ in the ratio $f/f_2$. Similarly, the corresponding figure for a 405-line video amplifier may be, say, 2·7 Mc/s.

*Step 4.* To find what value of $n$ leads to exact equalisation of the upper 'half-power' point.

The requirement here is that at the frequency $f_2$ the gain at H.F. shall *exactly* equal the gain at M.F. This involves a degree of compensation which lifts the response curve by +3 dB (i.e. to the 0 dB line) when $f=f_2$ (i.e. at the point on the $x$-axis where $f/f_2=1$).

Examination of Eq. (17) shows that the necessary condition for the gain at H.F. to equal the gain at M.F. is that:

$$1+n^2\left(\frac{\omega}{\omega_2}\right)^2 = \left\{1-n\left(\frac{\omega}{\omega_2}\right)^2\right\}^2+\left(\frac{\omega}{\omega_2}\right)^2$$

But this equality is, in the case cited, to occur when $\omega/\omega_2=1$,

$$\begin{aligned} \therefore\ 1+n^2 &= (1-n)^2+1 \\ 1+n^2 &= 1-2n+n^2+1 \\ 2n &= 1 \\ n &= 0{\cdot}5 \end{aligned}$$

As shown in Fig. 6.11, this value of $n$ leads to a slight hump in the response curve just below the frequency at which $\omega/\omega_2=1$.

# Index

HUTSON

Television Receiver Theory

1

ARNOLD